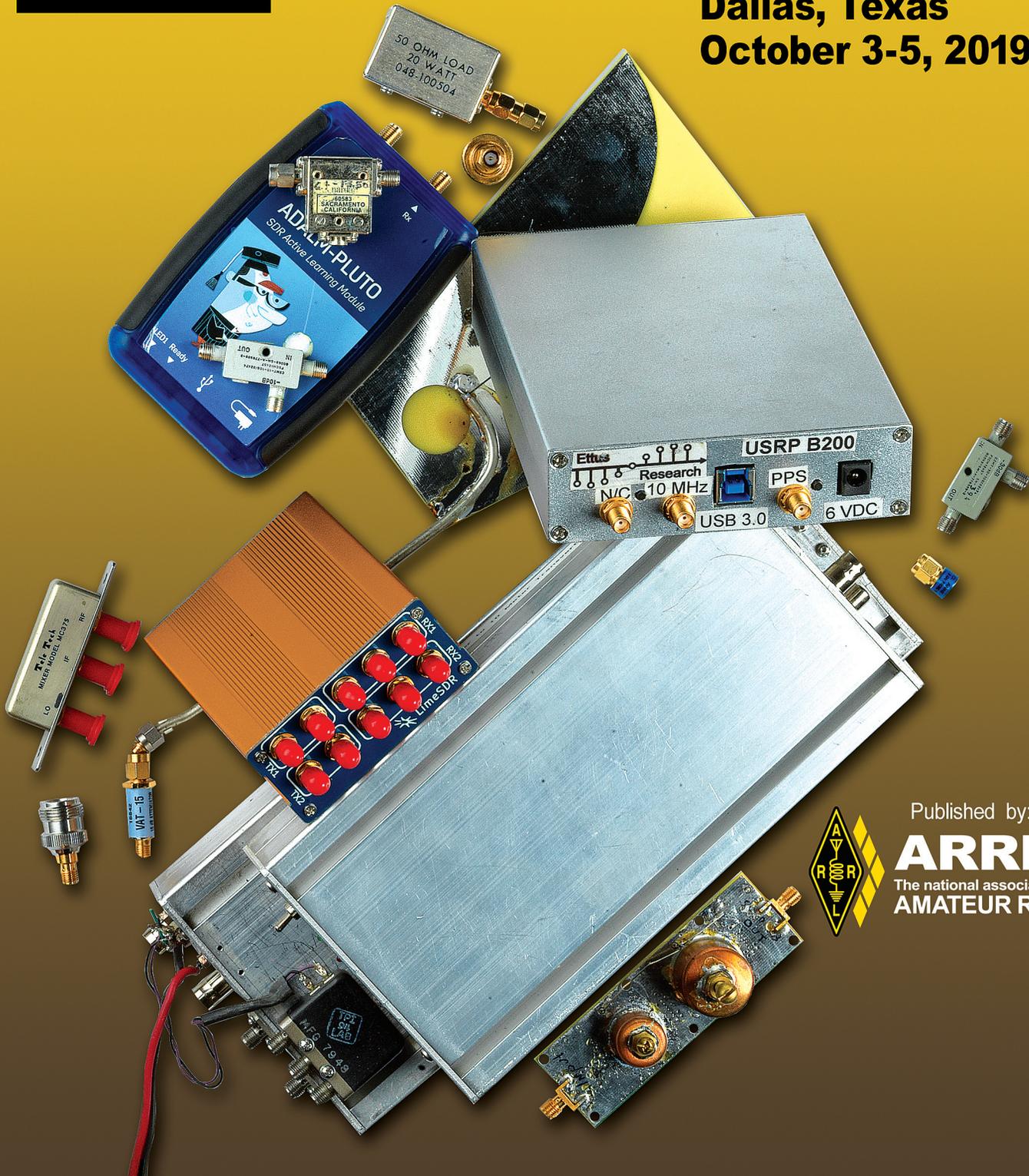


PROCEEDINGS OF

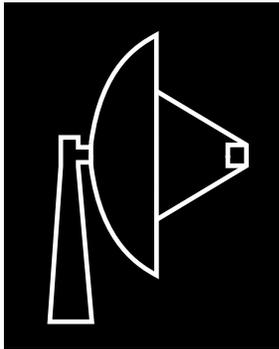
MICROWAVE UPDATE 2019

Dallas, Texas
October 3-5, 2019



Published by:

ARRL
The national association for
AMATEUR RADIO®



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On the cover:

The images on the cover were taken by N5BRG using a Hasselblad 500 cm camera and an HC Macro 120 II lens. Two Soft light boxes and reflectors were used to lite the scene. The resulting images have a resolution of 100 megapixels and the file size for the image is about 300 MB. The idea was to mix a combination of SDR radios and traditional microwave radio components, which is a recurring theme in today's amateur radio conferences and equipment.



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First Edition

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Welcome to Microwave Update 2019

Hosted by

North Texas Microwave Society

Hello everyone and welcome to the 37th edition of Microwave Update, MUD 2019. This conference is one of the technical conferences held every year to allow people devoted to the hobby to come together and exchange new ideas and learn new tricks. We believe this year will continue the tradition.

Initially the microwave radio was a transverter and a base band radio from a commercial manufacture. A few were complete home brew rigs. Today we are seeing more and more digital equipment with SDR radios being used for base band and complex integrated circuits providing compact reference oscillators with GPS slaved time references. Complex test equipment is coming down in cost with the proliferation of ICs either targeted for the application or some clever ham finds a way to use a communication IC for test equipment. There is no sign of this trend easing. We know a lot of you were blazing frontiers when you got started with microwave radio and keeping up with all the technology just keeps the juices flowing.

This year we decided to have a workshop on Gnu Radio with Tom McDermott, N5EG leading the workshop. Tom introduced the GNU Radio to MUD last year. This is a development package that can be used to build radios, beacons, and test equipment. GNU Radio has been around for many years but it is really just gaining momentum with advent of many low cost radios for interface. We hope you get a lot out of this workshop.

Many in this group are growing old and the sad truth of it is we are losing some real troupers like Bill Tynan, W3XO, Bob Gormley, WA5YWC, and Keith Pugh W5IU, Jim Johns, KA0IQT and others we will all miss. I am sure they are with us now and pushing us on. We wont forget the good times and QSOs we had with these guys and all the other SKs.

Please be sure to thank the speakers for taking the time and effort to share their knowledge and experience with us all. We hope you are inspired by these talks and go home and do something with the ideas. Please give some thought to participating in future MUD events and share your own experiences.

Finally we are working in the Dallas area to bring young people into our group. We are fortunate to have a few join NTMS recently. Please try hard to do the same thing in your area. Help them with equipment and support to get on the air. We are competing with cell phones and a fast moving society so we have to keep doing hi-tech interesting things to attract young people and maintain the old ones as well. We need this to support future MUDs and to maintain our frequencies.

If you are in Dallas for MUD 2019 get all you can out of the time. If you are reading this in the proceedings and did not make the trip be sure to consider attending Microwave Update in the future. There is nothing like being at MUD!

History of Microwave Update

By Al Ward W5LUA

2019 Revision

In 1985, Don Hilliard, WØPW, felt the need to organize a conference dedicated to microwave equipment design, construction, and operation. At the time of its conception, many microwave terrestrial and EME firsts were occurring on the microwave bands and it appeared that microwave needed a dedicated conference. Don held the first conference which he named “The 1296 and 2304 MHz Conference.” It was held at the Holiday Inn in Estes Park, Colorado. 66 people were in attendance. It sure seemed like Don was on the right track with his idea and he was right. In 1986, Don held the second conference which he rightfully named “Microwave Update 86.” 64 people were in attendance. The 1987 and 1988 “Microwave Update” conferences were again held in Estes Park, CO, and chaired by Don Hilliard.

After putting on 4 fine conferences in Colorado, Don decided to take a break from all of the work. Don turned over the responsibility of coordinating the event to the North Texas Microwave Society (NTMS).

In 1989, WB5LUA and WA5VJB of the NTMS hosted the 5th “Microwave Update” in Arlington, Texas where 94 people were in attendance. The 1990 “Microwave Update” was to go back to Colorado where Keith Ericson, KØKE and Don Lund, WAØIQN, were to head up the event. Unfortunately, Don Lund passed away during the year and Keith decided to postpone the 1990 Update. WB5LUA and WA5VJB of the NTMS hosted “Microwave Update” 91 in Arlington, Texas.

“Microwave Update” ’92 was held in Rochester, New York and sponsored by the Rochester VHF Group. The conference was chaired by Frank Pollino, K2OS and Dave Hallidy, K2DH (x KD5RO/2). “Microwave Update” ’93 was held in Atlanta, Georgia. The conference was organized by Jim Davey, WA8NLC, and assisted by Rick Campbell, KK7B and Charles Osborne, WD4MBK. “Microwave Update” ’94 was brought back to Estes Park, Colorado where it was chaired by Bill McCaa, KØRZ. Bill was assisted by Al Ward, WB5LUA, Jim Davey, WA8NLC, Jim Starkey, WØKJY, Phil Gabriel, AAØBR, and other local area amateurs. “Microwave Update” ’95 was brought back to Arlington, Texas and was chaired by Al Ward, WB5LUA and Kent Britain, WA5VJB of the NTMS. The 96 “Microwave Update” was held in Phoenix, Arizona and was chaired by Jim Vogler, WA7CJO. The ’97 “Microwave Update” was held in Sandusky, Ohio and sponsored by Tom Whitted, WA8WZG, with the assistance of Tony Emanuele, WA8RJF. The 1998 “Microwave Update” was held in Colorado under the guidance of Bill McCaa, KØRZ, and John Anderson, WD4MUO. The 1999 “Microwave Update” was held in Plano, Texas with Al Ward, W5LUA and Kent Britain, WA5VJB hosting the event.

The 2000 Microwave Update was held in the Philadelphia area with John Sortor, KB3XG, and Paul Drexler, W2PED hosting the event. The 2001 Microwave Update was hosted by Jim Moss, N9JIM and Will Jensby, WØEOM, in the Sunnyvale, California area. The 2002 conference was held in conjunction with the Eastern VHF/UHF Conference in Enfield, CT. The conference was hosted by Paul Wade, W1GHZ, Matt Reilly, KB1VC, Tom Williams, WA1MBA and Bruce Wood, N2LIV.

The 2003 conference moved across country to Seattle, WA where Rick Beatty, NU7Z and the PNWVHFS hosted the event. Rick's committee consisted of John, N7MWW as the co-chairman along with Jim, K7ND, Eric, N7EPD, Jim, W7DHC, Jimmy, K7NQ, and Lynn, N7CFO.

The 2004 conference was held in Dallas, Texas where Al Ward, W5LUA, Bob Gormley, WA5YWC, Kent Britain, WA5VJB, and the North Texas Microwave Society hosted the event.

The 2005 conference was held in Cerritos, CA. The event was hosted by Pat Coker, N6RMJ and Chip Angle, N6CA, along with the San Bernardino Microwave Society and the Western States Weak Signal Society.

The 2006 conference was held in Dayton, Ohio and was hosted by Tom Holmes, N8ZM, and Gerd Schrick, WB8IFM of the Midwest VHF/UHF Society.

The 2007 conference was held in Valley Forge, PA at the Dolce Valley Forge. The conference was hosted by Phil Theis, K3TUF, David Fleming, KB3HCL, Rick Rosen, K1DS, and Paul Drexler, W2PED of the Mt. Airy VHF Radio Club.

The 2008 conference was hosted by Donn Baker, WA2VOI, Barry Malowanchuk, VE4MA, Jon Platt, WØZQ, Bruce Richardson, W9FZ, and Bob Wesslund, WØAUS, of the Northern Lights Radio Society, and was held in Bloomington, MN.

The 2009 conference was held in Dallas, Texas and was hosted by Steve Hicks, N5AC, Al Ward W5LUA, Bob Gormley, WA5YWC, and Kent Britain WA5VJB of the North Texas Microwave Society. In 2009, The Don Hilliard Technical Achievement Award was created in honor of our founding father Don Hilliard, WØPW. The first recipient was Paul Wade, W1GHZ in recognition of his many years of service to the amateur microwave community.

The 2010 conference was hosted by the San Bernardino Microwave Society in Southern California and was chaired by Pat Coker, N6RMJ and Doug Miller, K6JEY. The conference was dedicated to Dick Kolby, K6HIJ who recently became a silent key. Dick was posthumously awarded the Don Hilliard Technical Achievement Award for his many years of dedication to the VHF and microwave community.

The 2011 conference was held in Enfield, CT, hosted by the North East Weak Signal Group. Bruce Wood, N2LIV (SK) was the conference chairman and Paul Wade, W1GHZ was the program chairman.

The 2012 conference was hosted by the 50 MHz and Up Group of Northern California in Santa Clara, CA. Jim Moss, N9JIM and Will Jensby, WØEOM, were the conference co-chairs.

The 2013 conference was hosted by Jeff Kruth, WA3ZKR and the Morehead State University at the Space Science Center at Morehead State University in Morehead, KY

The 2014 conference was hosted by the Rochester VHF Group and organized by Dave Hallidy, K2DH, Frank Pollino, K2OS, and Bill Rogers, K2TER, and was held at the Marriot Airport Hotel in Rochester, NY.

In 2015, the 30th Microwave Update conference was hosted by the San Bernardino Microwave Society as it celebrated its own 60th year and with support from the San Diego Microwave Group. Chaired by Pat Coker, N6RMJ, the conference was held in San Diego, CA, and included presentations on three world distance records set during the past year.

The 2016 Microwave Update Conference was hosted by the St. Louis Area Microwave Society and was held at the Holiday Inn Airport West in Earth City, MO, near St. Louis. The organizers were Bob Sluder, NØIS, Mel Whitten, KØFPX, and Jim Richardson, N5MU.

The 2017 Microwave Update Conference was hosted by the 50 MHz and Up Group of Northern California. It was held at the Biltmore Hotel in Santa Clara, CA. Jim Moss, N9JIM and Mike Lavelle, K6ML, were the conference co-chairs.

Microwave Update 2018 was held in Fairborn, Ohio, a suburb of Dayton, and was hosted by the Midwest VHF/UHF Society, an active group of hams mainly in southwest Ohio, but with members all over the US, Canada, and beyond. A highlight of the weekend was a tour of the Voice of America Museum in Mason, Ohio, and the 10 GHz EME station setup at the VOA site.

Anyone interested in sponsoring a future conference may contact the selection committee: Kent Britain, WA5VJB, Barry Malowanchuck, VE4MA, and Tony Emanuele, K8ZR (ex WA8RJF).

Submitted by Al Ward W5LUA August 20, 2019

Microwave Update 2019 Committee

Chairman: Al Ward, W5LUA, and Bob Stricklin, N5BRG

Banquet Prizes: Eric Haskell, KC4YOE

Registration: Jim Hudson, WA5JAT, and Ross Pounders, K5ZSJ

Treasurer: Wes Atchison, WA5TKU

Secretary: Eric Haskell, KC4YOE

Web Site: Dave McCoy, N5RIJ

Audio Visual: Mike Watts, WY6K, and Bob Stricklin, N5BRG

Photographer: Al Webb, W5RLG

Proceedings: Kent Britain, WA5VJB, Al Ward, W5LUA, and Bob Stricklin, N5BRG

Digital Copies: Jim Hudson, WA5JAT and Bob Stricklin, N5BRG

Hospitality Room: DEMI and Directive Systems

Surplus Tours: Kent Britain, WA5VJB

Auction: Kent Britain, WA5VJB

GNU Radio Workshop: Tom McDermott, N5EG, Bob Stricklin, N5BRG, Jenner Lochridge, KK6RUM, and Eric Haskell, KC4YOE

Antenna Range: Kent Britain, WA5VJB

Noise Figure Workshop: Al Ward, W5LUA, Tony Emanuele, K8ZR, and Doug Carlson, ADØCX

Phase Noise Workshop: Chuck Clarke, AF8Z, and Eric Haskell, KC4YOE

Family Program: Emily Ward and Bob Stricklin, N5BRG

Airport Transportation: Bob Stricklin, N5BRG

Vendor Tables: Al Ward, W5LUA

The Don Hilliard Technical Achievement Award

Don Hilliard, W0PW, (exWØEYE) an early VHF pioneer was involved with the formation of the Central States VHF Society back in 1967. The Central States VHF Society was and still is very instrumental in promoting VHF and above activity. Back in 1985, Don realized that there was a considerable thrust in new microwave technology above 902 MHz. As a result, Don felt the need to have a conference devoted to the higher frequencies. The conference would be devoted to microwave equipment design, construction, and operation. At the time of its conception, many microwave terrestrial and EME firsts were occurring on the microwave bands and it truly appeared that microwave needed a dedicated conference. Don organized the first conference which he named "The 1296 and 2304 MHz Conference". It was held at the Holiday Inn in Estes Park, Colorado. 66 people were in attendance. It sure seemed like Don was on the right track with his idea and he was right. In 1986, Don held the second conference which he rightfully named "Microwave Update 86". 64 people were in attendance. The 1987 and 1988 "Microwave Update" conferences were again held in Estes Park, Co. and chaired by Don Hilliard.

After putting on 4 fine conferences in Colorado, Don decided to take a break from all of the work. Don turned over the responsibility of coordinating the event to the North Texas Microwave Society (NTMS). The rest is history. With the exception of one year where one of the organizers, Don Lund passed away, Microwave Update has been held every year. To this date including the 2019 conference being held in Lewisville, Texas, Microwave Update has been hosted 34 times. The conference has been successfully organized and run by various local VHF and microwave clubs and groups around the US.

In tribute to Don Hilliard and his tremendous contributions to VHF and microwave technology and for appreciation of his forward looking into the fascinating world of "microwaves," the North Texas Microwave Society on behalf of Microwave Update created "The Don Hilliard Technical Achievement Award" presented each year to an amateur radio operator who has made significant contributions to amateur microwave operation and technology. The NTMS proposes that this award be presented to a deserving amateur each year by each sponsoring organization.

The first Don Hilliard award was presented to Paul Wade, W1GHZ in 2009 for his many papers, projects and On-Line Antenna book. The 2010 award went to Will Jensby, WØEOM for his pioneering microwave and SHF work. The 2011 award went to Brian Justin, WA1ZMS, the first (and only, to date) holder of the Worked All Bands award, for his record-breaking work on the millimeter wave bands and for his outstanding papers on many subjects. The 2012 award went to Goran Popovic, AD6IW for his dedicated work within the hobby. In 2013, the award went to Tom Williams, WA1MBA for his millimeter wave work, the 78 GHz LNA project, and numerous papers and presentations. In 2014, the award went to Al Ward, W5LUA, The 2015 award was presented to Ed Munn, W6OYJ. The 2016 award went to Barry Malowanchuk, VE4MA. The 2017 award went to Sergei Zhatyayev, RW3BP for his pioneering work on 47 and 76 GHz EME. The 2018 award went to Michelle Thompson, W5NYV for her work in digital signal processing, software defined radio and GNU radio.

Respectfully submitted,

Al Ward, W5LUA
Kent Britain, WA5VJB
September 3, 2019

Author Biographies

Tom Apel / K5TRA - Biography

Tom Apel is an electrical engineer. He is currently serving as vice president of the Roadrunners Microwave Group, an ARRL affiliated club. Tom chaired the technical program for the 2014 Central States VHF Society conference. In 2010 he retired from Triquint Semiconductor as Senior Engineering Fellow where he managed advanced component development. Mr. Apel has 33 years in microwave and RF component design at VHF through Ka band. He developed the first 6-18 GHz 2W power amplifier MMIC to achieve volume production. More recently, his work has resulted in many power amplifier products for handset applications. During his career he was responsible for 35 US patents. He earned a BS Physics and BS Mathematics from Loras College, and MSEE from University of Wisconsin, Madison. Tom was first licensed in 1963 and has been home brewing since then.

Alan Devlin, VK3XPD.

In March of 2019. I reached the milestone age of 68 Years and “some” months!

Retirement living in Melbourne, Australia has been quite enjoyable with plenty to do.

As for my past career... I spent most of my working life employed by Australia’s leading TELCO – Telstra, formerly known as the Post Master General’s Department (PMG). After leaving Telstra in 1998, I started and operated a very successful one-man small Business – RF RESALE, a WEB based Buy, Sell and Repair Test Equipment Business. After a few years, I got tired of dealing with the Australian Taxman and closed this Business down. That was more than 10 years ago, and I’ve not regretted a thing!

I have no formal Engineering qualifications. All of my knowledge and experience has been developed by getting my hands dirty, pulling things apart, modifying RF hardware to make it work where it was never intended and learning from my Peers.

My first exposure to the wonderful world of RF was when at the age of 12, when I built my first Crystal Set using liquid Solder from a tube.

I was a late entrant into Amateur Radio being first licensed as VK3XPD in the late 1980’s. Along the way I have aligned myself with other likeminded RF Hobbyists and together we have built up numerous items of RF gear to then use to Set or Extend many Australian Microwave and Millimetre Wave Distance Records up to 324 GHz.

My current Amateur Radio focus continues to be in the Millimetre Waves.

It’s been a wonderful Amateur Radio journey and I hope it will continue for many more years to come.

Warren Ferber, WFØT

Warren was first licensed in 1976. Since then he has held every class of license. He was intrigued by the Technician license due to its focus on the VHF and above allocations and spent several years chasing states via meteors on 6 meters.

Like a moth, he was drawn to the light. In the 90's when laser pointers first came out, he quickly seized on the opportunity and modulated one for a beacon. He experimented with the detectors and optical chips available and spent many a night bouncing lasers off various buildings, signs, and objects listening for the returns on his receiver.

Warren is an avid golfer, he plays on the NBC Golf Channel Amateur Tour, enjoys running, camping, photography and has an understanding Wife who doesn't mind standing outside in the cold running a light transmitter while he tests out a new receiver a mile away. Now he wants to help other hams see the light, or in the case of Infrared, hear the light.

Rob Jahnke, KØXL

Rob began his radio interests as a SWL'er around age ten with an AM/Shortwave console radio purchased at auction. He was fascinated with being able to listen to AM DX like KOA (Denver), WLS & WGN (Chicago), and WSM (Nashville) and many others from his home in Wisconsin. As a freshman at the University of Wisconsin, he discovered amateur radio. He received his Novice license and built his first CW transmitter and superhetrodyne receiver. Several of his contacts were on 80 meters using the rain gutters outside his dorm room as an antenna. He now holds an Extra class license.

His early working years were at IBM where he became interested in microprocessors and PC's and how to apply them to amateur radio projects. One project was a remote base 2 meter radio operated with touch tones over a phone line. It used an Intel 8048 single chip microprocessor and used 1023 of the 1024 bytes available.

Rob's interests lie more with experimenting than operating. He views operating to be a way to verify his experiments. Optical communications have long been a fascination for Rob and when he saw Warren Ferber's presentation at Aurora on using a single chip microprocessor (ATTiny85) as a controller/modulator for optical communications with LED's he quickly joined him in this pursuit.

Rob holds a Bachelor's Degree in Mechanical Engineering and a Master's Degree in Electrical Engineering, both from the University of Wisconsin – Madison.

Ben Lowe, K4QF/K4VOW

Ben Lowe, K4QF/K4VOW, was licensed as KN4VOW in 1958 at the age of 14. He holds an MS degree from Southern Methodist University in Dallas and a BS degree from The University of Tennessee at Knoxville, both in Electrical Engineering. He is a registered Professional Engineer in the States of Alabama and Texas. Before retirement, Ben designed and implemented microwave systems, assisted the military with GPS applications and wireless architecture development, and conducted RF environmental analysis and measurements for Engineering Systems & Planning (ESP) of Huntsville, Alabama. Previous employment included: Collins Radio Co., Texas Instruments Co., Stanford Research Institute, The BDM Corp., and Science Applications Int'l Corp.

In addition to Amateur Radio, he was a licensed pilot, had an advanced SCUBA certification, and was is an avid bicyclist. His primary ham radio interests are in weak signal VHF/UHF/Microwave operation. He has written numerous articles for *QST*, *CQ*, *Ham Radio*, and *The ARRL Handbook*. Previous call signs included WA5UVM, ZL1AQC, G5EPZ, K4QF/HB/DL/HBØ and ZF2BL. You can reach Ben at 848 County Rd #138, Scottsboro, AL 35768, or at k4qf@arrrl.net.

Skip MacAulay, VE6BGT

Skip was licensed in 1976 and started off building as much of his amateur equipment as possible. Building many types of HF, VHF and UHF amplifiers and equipment he finally got to his ultimate goal around 2008 and made his first EME contact on 23 centimeters. As the saying goes you always want a bigger antenna, he went from a 10 foot TVRO dish to an expanded 12 foot and then finally built a 21 foot mesh dish. Continuing with his passion of home brewing he is now on 23, 13, 9, and 6 centimeter

moonbounce. Skip was employed for the last 30 years mostly in the Oilfield electronics field in Western Canada and offshore the Canadian east coast on such platforms as Hibernia.

Barry Malowanchuk, VE4MA

Barry was active on 432 in 1966 and 10 GHz in 1968 using his father's call VE4MA and was first licensed as VE4MA in 1975. Barry has been active on moonbounce (EME) since 1974, and has operated EME on all bands from 432 MHz -47 GHz, participated in the first ever 24 GHz QSO (WFP Front Page article), and has been involved with the first 78 GHz EME experiments.

Barry has authored and presented many amateur conference papers on EME, dish feedhorns, solid state & vacuum tube power amplifiers, and low noise amplifiers. Barry has received the Central States VHF Society John Chambers Award in 2000 & again in 2008, the ARRL's Microwave Development Award in 2003, the Northern Lights Radio Society's Wesslund Award in 2008, and the Microwave Update Don Hilliard Award in 2016.

Tom McDermott, N5EG is retired and has been licensed 50 years. He is a member of TAPR, and a life member of ARRL. Tom has written numerous articles in *Ham Radio Magazine* covering TEXNET, and *QEX* covering topics from ionospheric measurements to gnuradio.

He has written one textbook "Wireless Digital Communications: Design and Theory."

Professionally, Tom worked in the development of fiber optics communications and switching systems in roles ranging from ASIC design to CTO, and was board member of several non-profit educational organizations. He is a Senior Member of the IEEE, and previously was a voting member of the IEEE Ethernet 802.3 working group. Tom holds 13 patents.

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Greg McIntire, AA5C

Greg McIntire, AA5C has a BSEE from Iowa State University and a MSEE from The University of Texas at Austin. He is a radar systems engineer and a registered professional engineer in the state of Texas (inactive). Greg retired from Raytheon (Texas Instruments) in 2006 as a senior engineering fellow and L3 Mustang Technology in 2017 as an engineering fellow. He is operational on all bands through 47 GHz and enjoys building, operating, and writing with articles published in *QST*, the **CSVHF Proceedings**, and the *Microwave Update Proceedings*. Greg was the 1998 recipient of the Central States VHF Society John T. Chambers Memorial Award.

Doug Millar, K6JEY

Doug Millar was first licensed as KN6JEY in 1957. He has mainly been interested in VHF and up communications. He currently is active from 1.2 GHz to 122 GHz. He has been active on EME on 144,

432, 1296 and is almost done with a 10 GHz small dish EME rig. He has been an ARRL Technical Advisor in calibration and test equipment and wrote the 26th chapter of the *ARRL handbook* (1994-2014). He was awarded an EdD. in 2005 and has been a teacher/professor for 50+ years.

Rex Moncur, VK7MO

Rex Moncur, VK7MO, was first licensed in 1960 and focussed on 144, 432 and 1296 MHz in the days of AM with 832s and 2C39s. He gained Australian records of just 136 miles on 432 and 22 miles on 1296 MHz. Rex graduated in Communication Engineering from the Royal Melbourne Institute of Technology in 1962 and took up employment with the Australian Bureau of Meteorology on the development, testing and procurement of telemetry systems including radiosondes and rocketsondes. In 1973 Rex moved to administrative roles in the head office of the Department of Science in Canberra and ham radio took a back seat. Roles ranged from industrial relations and personnel policy, Secretary of the Australian Research Grants Committee, to head of Resource Management of the Department of Science. In 1988 Rex was appointed as Director of the Australian Antarctic Division in Hobart, Tasmania responsible for the administration of the Australian Antarctic Territory including operational support for the Antarctic program and the management of the Australian Antarctic Science program. He represented Australia in negotiating the Madrid Protocol to the Antarctic Treaty for the Protection of the Antarctic Environment. Rex retired in 1999 and returned to his interest in Amateur Radio and portable operations. Initially on 144 MHz with FSK441 and then moved to higher frequencies with JT65 and up to 10 and 24 GHz with JT4 and QRA64 and also optical cloud-bounce. Rex currently holds the 10 GHz terrestrial World record and the 10 and 24 GHz EME World records.

Dave Robinson, WW2R, G4FRE

Dave was first licensed in 1974 as G8JMO, upgraded to G4FRE in 1976 and WG3I in 1989, recovering his mis-issued WW2R call from the FCC in 1996. He received a BSc in Electronic Engineering from the University of Wales in 1979. At University (GW3UCB/P) he was introduced to VHF/UHF contesting, which continued when he joined British Telecom (G4BPO/P) with the Sheppey Contest group (G4NXO/P) then after coming to the USA with the Mt Greylock Expeditionary Force (W2SZ/1). Operated from 2005 to 2012 on 144-5760MHz EME then returned to chasing DX on the microwave bands. He currently works for Element Materials Technology, based in Minneapolis as an EMC test engineer.

Bob Stricklin, N5BRG

Bob was first licensed in 1969 and for the next 10 years spent more time operating a radio than he does now. After getting a BS in Physics he started working in the electronics industry designing and building optical and magnetic sensors. This included sensors for computer paper tape and punch card readers. This along with college classes in Fortran IV and Cobol introduced him to computer systems and digital electronics. Then there was a quest to get a personal computer (Apple II) and constantly upgrade to the newest fastest he could afford. He attended early meetings of the Dallas area computer users group and talked about the IMSAI 8080. Later, it was an S100 bus computer and today it's a matter of managing all the computers in the house including Linux, Mac, and Windows. Along the way Bob joined AMSAT and operated some of the early satellites then he got involved with building amateur satellites and contributed by building some of the space hardware. He also got involved in TAPR at its start and built one of the first 30 prototype TNCs to interface with radios for amateur use and put this on the air in Dallas. In 1993 he designed and built a DSP based radio interface called DSP93 and this was sold as a kit by AMSAT/TAPR and used for some of the early entry into SDR by amateurs. He has been active at various levels in North Texas Microwave Society. Bob has given

presentations at many hamfests the ARRL Digital Communication Conference and local ham clubs. Before retiring from Austria Micro Systems in his professional career he worked on sensor applications for medical, automotive, space applications, fiber optic communications and cell phones. These sensors grew in complexity over the years to CMOS ICs with digital state machines requiring Verilog firmware and software drivers. He had lots of help along the way from good friends and his wife Lillian. He enjoys giving back to young and old. Bob is life member of the IEEE.

Paul Wade, W1GHZ

Paul Wade W1GHZ, previously N1BWT and WA2ZZF, has been licensed since 1962 and has never made a contact below 50 MHz. He has been a microwave experimenter for years and published numerous articles as well as writing the “Microwavelengths” column in *QST*. He is active in the Vermont 10GHz group and is past President of the North East Signal Group. An ongoing project is the “W1GHZ Microwave Antenna Book” – online at www.w1ghz.org.

In 1997, he was honored to be named by the Central States VHF Society as the recipient of the Chambers Award. He has been honored by the ARRL with the 2000 Microwave Development Award, in 2001 with the Thomas Kirby Eastern VHF/UHF Society Award, and in 2009 by Microwave Update and the North Texas Microwave Society with the Don Hilliard Award for Technical Contributions to the Microwave Community.

Paul is a Life Member of the IEEE. After a long career in electrical engineering, including microwaves, computer hardware, and signal integrity, he and Beth, N1SAI, are now happily retired in Vermont with a new puppy named Hannah. He was also a ski instructor for a time, and now enjoys skiing on a new bionic knee, and skijoring with Hannah.

Al Ward, W5LUA

Al graduated with a BSEE from the University of Illinois in 1973. He was a System/Circuit Designer at Texas Instruments from 1973 to 1987 working on the Shrike and HARM missile programs plus the IFF program. From 1987 until May 2012, he was a Semiconductor Applications Engineer with Avantek, Hewlett Packard, Agilent Technologies, Avago Technologies and now Broadcom. His specialty is the design of low noise amplifiers. He has been retired since 2012. Al was first licensed as WN9QZE in 1965 and presently holds the Amateur Extra Class ticket. Al operates all amateur frequencies from .474 MHz through 122 GHz. Al has WAS on 50, 144, 220, 432 MHz and 1296 MHz. and WAC on 50, 144, 432, 1296, 2304, 3400, 5760 and 10368 MHz. and has worked 36 states on 2304 MHz. Al recently received DXCC #7 on 1296 MHz which gives him DXCC on 11 bands. Al has completed EME QSO's on all bands, six meters through 47 GHz and has received RW3BPs 77 GHz signal via the moon. Al currently holds the US distance record on 10 GHz terrestrial with K0VXM at 1609 km. Al was instrumental in the formation of the North Texas Microwave Society and was the founding President. Al has received the Central States VHF Society John Chambers Award, the Central States VHF Society Mel Wilson Award, the Southeastern VHF Society K4UHF Award and was the recipient of the 1997 Dayton Hamvention Technical Excellence Award. Al has also received the ARRL's 1999 Microwave Development Award and the Microwave Update sponsored Don Hilliard Technical Achievement Award.

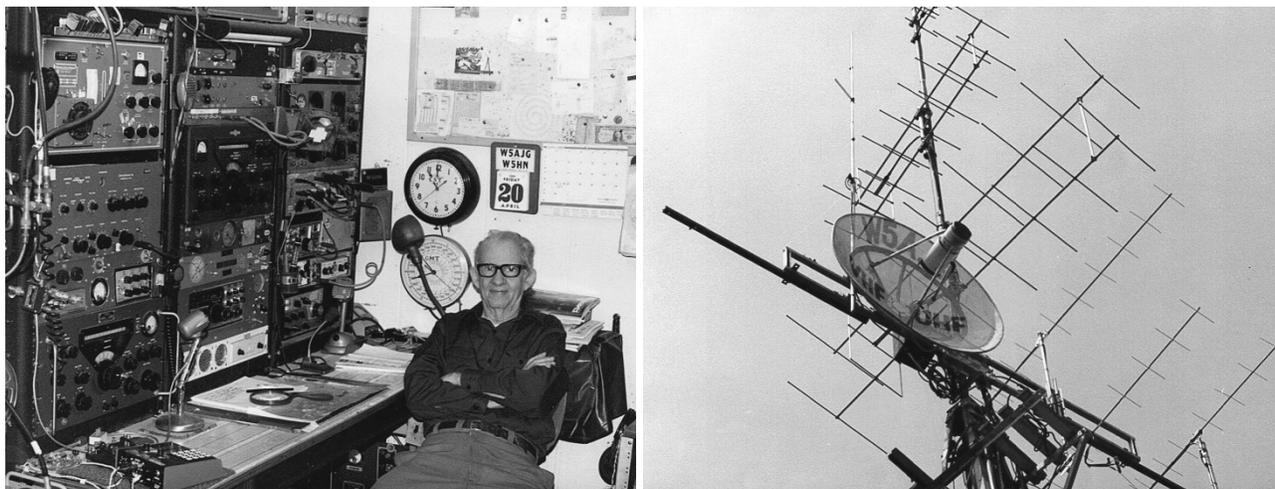
Thomas D. Williams, Ph.D., WA1MBA

Tom Williams was first licensed at the age of 14 as K1URO. He earned his Ph.D. in Computer Science from the University of Massachusetts in 1981, and has founded several product companies in the areas of millimeter wave imaging and image understanding. He lives in Massachusetts, and is

presently consulting for a drone company in Kansas. Tom has decades of experience designing and implementing systems which operate in the EHF spectrum, and in the 1990's achieved the first QSOs on the 145 and 120 GHz bands. Tom is a member of the North East Weak Signal group, has had the pleasure of creating and administering the Microwave email reflector, and has built a number of 78 GHz LNAs for the amateur community. He has numerous technical publications and six patents. His favorite annual activities are the 10GHz and up contest, Red Sox Spring Training, and various ham radio meetings including the Microwave Update Conference.

The North Texas Microwave Society

Club Call W5HN



Leroy May

5AJG W5AJG W5HN

Leroy was the Patriarch of VHF/UHF/Microwave activity in the North Texas area. Leroy was the last station in Dallas running Spark. As he tells the story about one afternoon in 1923, he came out of his High School and nearly half of the hams in Dallas were waiting for him. He was invited across the street for a Root Beer. Together they had gotten a tube, a porcelain socket, and someone willing to rewind his spark transformer. In short, all the parts needed for Leroy to build a CW transmitter and get him off spark!

His interest in VHF started early. Leroy only needed 6 more states for WAS on 5 Meters, when WWII closed down the ham bands. After the war Leroy restarted on 6 Meters completing his 48 state WAS in 1950. Seems it took quite a bit of arm twisting to get one of the Dallas hams to make a DX pedition to Arkansas to give Leroy his last state.

In the early 50's Leroy built a 20A Phasing SSB transmitter, and a 50 MHz transmit converter modified from a WWII noise jammer and soon he was on the air. Only 3 watts, but it was SSB! A few years later a pair of 4X150's rounded out the station. I always liked his 5 element 6 Meter beam arrangement. The beam was at about 30 feet on an old 4"x4" wood beam with a Model T Ford rear bearing at the base. His rotator was a rope wrapped about the wood beam. The rope ends came back into his shack. He would hear a weak station, open the window, pull the ropes back and forth until the

signal peaked, then closed the window on the ropes to hold the antenna in position. True ‘Armstrong’ rotator.

By the end of the 1950’s Leroy had added 144 MHz and 220 MHz SSB stations. Leroy was the only 220 MHz station in the 5th call district for nearly 20 years, and the only one in North Texas until WB5LUA and WA5VJB built stations in 1980.

Leroy built his 432 MHz and 1296 MHz SSB stations in the 1960’s. I first heard of Leroy in 1969 and was told of his morning 1296 MHz schedules with W5LDV. Chic was south of Houston, and a nearly 300 mile path, and to think you could have regular QSO’s over that distance on that frequency was amazing to me. Leroy was one of the few people to acquire, modify, and get on the air a 6 tube UPX4 2C39 amp for several hundred watts output. Big power for that time.

Leroy also knew Sam Harris, W1FZ and the founder of Microwave Associates. Leroy ended up with one of Sam’s commercial Para-amps. State of the art technology for it’s day. W5LUA and I made several GaAs FET preamps for Leroy, but none of them ever worked as well as that Microwave Associates Para-Amp. (Note A Parametric Amplifier excites a Varactor Diode with microwave energy and uses that energy to amplify a lower frequency signal with high gain and very low noise.) It was after Leroy passed away just shy of his 90th birthday that I had that Para-Amp on my test bench and figured out what was happening. That Leroy had ever gotten it working on 1296 MHz was amazing. The amplifier was obviously built for 1.0-1.1 GHz L-Band use. In that frequency range you could dump just about any X-Band energy into the ‘Pump’ port and it amplified. As you moved the amp up in frequency the pump got more and more critical. At 1296 MHz the pump had to be in a 50 MHz window just above 12 GHz and in a narrow power range. Now it had 27 dB of gain and a 2.5 dB NF. Sure, we can beat that with a GaAs FET, but Para-Amps are a regenerative amplifier with a narrow bandwidth. Wideband down at 1 GHz, but only 2 MHz wide at 1296 MHz. Acting like a very narrow filter, all of Leroy’s 1296 MHz images and LO spurs were gone!

In the late 50’s, Wayne Green was a cameraman for the Dumont TV station in Dallas. His publishing empire consisted of a Mimeographed newsletter on teletypes. Wayne would come by Leroy’s QTH on Sunday afternoon with a 6 pack of beer. Now Leroy was a teetotaler, and was privately offended by this but let Wayne drink on his back porch. As 73 Magazine got started Wayne was always tapping Leroy for feature articles. In one issue of 73 Leroy had 5 articles.

Terry, W5ETG was working with the MARS organization at the Travis Air Base in San Antonio. Terry was a bit jealous of Leroy because it seemed like every time a new piece of Air Force equipment hit the MARS pile, the first one went to Leroy. But next month, there in the MARS newsletter was an article on converting to a ham band. Terry later agreed he was the best person for the first one.

Lastly, I want to mention Leroy’s 2304 MHz station. His 2304 MHz Ratrace Mixer was even a feature article in a 1970’s issue of QST. The transmitter chain was some old 1.8 GHz Microwave Link transmitters that had been used between Dallas and Houston. Not a lot of gain at 2.3 GHz, so he needed 3 to get up to the 10 watt level.



But now for the rest of one of my favorite stories. In 1979, using some parts that didn't meet spec from a work program, WB5LUA built a 2304 MHz station. Leroy dusted off some stuff he hadn't used in nearly a decade and gave Al his first QSO on 2304 MHz.

The North Texas Microwave Society honors Leroy by adopting W5HN as our Club Call.

Amazing Ham! Kent WA5VJB Sept 6, 2019

Microwave Update 2019 Schedule

Microwave Update 2019 Schedule		
Thursday October 3		
8AM to 2:30PM	Surplus Tour	Meet in lobby
2PM to 6PM	Registration	Piazza
3PM to 6PM	Gnu Radio Workshop	Tuscany Ballroom IV
6PM to 8PM	Dinner – on your own	
8PM to Midnight	Hospitality Suite	
Friday October 4		
7:30AM	Registration	Piazza
8AM to Noon	Noise Figure / Phase Noise	Colosseum
8AM to Noon	Ant Gain Measuring	Parking Lot
9AM	Family Program	Meet in lobby
Noon to 1:15PM	Lunch Seminar - WA1MBA	RainScatter
1:15PM - 2PM	WF0T/K0XL	Adventures at 630-850 nm
2PM to 2:45PM	VK7MO	Small Dish Portable EME
2:45PM to 3PM	Break	
3PM to 3:30PM	AA5C	VHF/UHF/Microwave Hi-Power LDMOS Power Amplifiers
3:30PM to 4PM	VE4MA	WR-28 Waveguide on 47GHz
4PM to 4:30PM	K6JEY	Using Ku Band LNBS on 10GHz
4:30PM to 5PM	K4QF	Rambo Roving for Microwaves
5PM to 7PM	Dinner - on your own	
7PM to 10PM	Flea market	Colosseum
7PM to Midnight	Hospitality Room	
Saturday October 5		
7:30AM	Registration	Piazza
8AM to 8:30AM	WA5VJB	Advanced Yagi Techniques
8:30AM to 9AM	N5BRG	A method to build your own Waveguide to Coax Transition

9AM	Family Program	Meet in lobby
9AM to 9:30AM	K8ZR	The microwave & millimeter wave bands and the regulatory challenges ahead
9:30AM to 10:00 AM	VE6BGT	6 cm Feed Experiments
10:00AM to 10:15AM	Break	
10:15AM to 11:15AM	K5TRA	Impedance matching techniques for RF and microwave circuit design
11:15AM to Noon	VK7MO	10 and 24 GHz Terrestrial DX
Noon to 1PM	Lunch Seminar – VE4MA	mmWave Fun in the Arizona Sun
1PM to 1:45PM	W1GHZ	Waveguide cross-guide directional couplers
1:45PM to 2:15PM	AA5C	Arduino Controllers for the Microwave Operator
2:15PM to 3PM	N5PYK	Case studies in microwave propagation by a NWS meteorologist
3PM to 3:15PM	Break & Auction	
3:15PM to 4PM	WW2R	Practical applications of the Lime SDR mini
4PM to 4:30PM	WA5TNY	Using the Texas Instruments LMX2594 synthesizer for microwave local oscillators
4:30PM to 5PM	WA5TNY	A coaxial waveguide dual band feed for S and X band
6PM	Cash Bar	
6:30PM	Banquet with after dinner speaker Rex Moncur VK7MO	
9PM to Midnight	Hospitality Suite	
Rev September 3, 2019		

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Rain Scatter, SHF and EHF

Tom Williams, WA1MBA – Microwave Update 2019

In 1999 I presented a paper and talk at Microwave Update on Rain Scatter. When asked about a topic for MUD 2019, I decided I could probably dust off a twenty year old paper, update it and present it anew. After all, many folk didn't pay attention the first time, may have forgotten it, and many would have not attended that conference 20 years ago. Because the EHF bands had changed in the intervening years, at least I had to execute some simulations to update the bands. That led me to discover that the original data were based on incorrect assumptions, and that all the polar plots I had presented two decades ago were at least slightly wrong. So, it was time to update all of those data. Fortunately, all of the basics were right, along with a few of the diagrams.

When we look up on a clear day we see a blue sky with white clouds. At sunset we might enjoy hues from red through yellow. We can take pictures through a polarizing filter to darken the sky to a rich deep blue. Look down at a soap bubble or a film of oil in a puddle and pretty colors appear to swim around. We make over-the-horizon contacts on ham bands all the time (see Figure 1). What do all these events have in common? Scattering.



Figure 1. Over-the-horizon contacts made by radio amateurs usually involve scattering.

The word “scattering” we can mean a variety of things – but generally scattering means that things go in a variety of directions - landing in different places often not in the direction thrown. We say “scattered to the four winds”, or “the population from that country is scattered across the globe”, or “Tom is scatter-brained”. This general sense of scattering is where the specific term that we use comes from because electromagnetic waves, when scattered by their interactions with matter, go in new directions, arriving in different places than they were originally directed, sometimes preserving polarization and modulation intelligence, sometimes not.

1 What is RF scattering?

The scattering of electromagnetic energy by matter is a complex subject, but there are a few relatively easy to understand principles that account for it. Fundamentally, the interaction between any EM wave and any matter is an example of scattering. This includes reflection and refraction. Because volumes have been written on the subject, and volumes more are yet to be written and understood (and because I am not really an expert), I can only cover very basic field-matter interaction and scattering in this report. Some simplifications are made here in order to understand the interactions pertinent to scattering in amateur radio communications. I will focus on rain as a scattering medium.

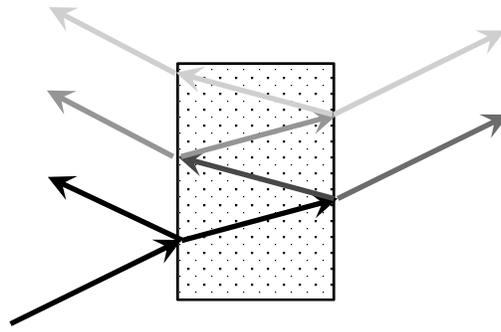


Figure 2. A simple RF interaction with a dielectric slab teaches some of the principles of general EM scattering.

Early attempts to understand scattering were unable to explain some observations. Those understandings were originally based on interactions with slabs of matter. For example, think of light hitting a clear thick glass plate, which reflects some waves off the backside (where the wave impinges on the object in this simplified diagram Figure 2). Some of the energy continues through the object, and the energy is again split back into the object (the glass) and some of the energy comes out the other side. Internal reflections continue with some “leaking out” on every bounce. In our example the glass has no absorption, but if it did, some of the internal energy would be converted to heat, so that all the emissions will sum to less than the energy entering. Although this explanation accounts for interactions with large slabs, it did not account for the scattering of light in many directions by particles. Also, the polarization that is exhibited by a group of particles is very different from what a slab-based explanation predicts. This and other problems led early researchers to concentrate efforts in unraveling the nature of wave/particle interaction, and that has led to a more thorough understanding of scattering.

A good way to think of scattering by material is to consider that the EM wave does at least three things. The first, like with a slab, is reflection. The interaction at the border between the transmitting medium (air) and the material (a dielectric like water or glass) is easily determined by the angle of the entry relative to the material surface, the polarization of the wave and the difference between the dielectric properties of the medium and the particle. Ignoring polarization effects, the reflection angle is the same as the incoming angle, and the power is split between the reflection and that which enters into the particle. The second thing a wave does, again like with a slab, is refraction. The wave propagates inside the particle at some angle, again determined by the entry angle, the relative dielectric properties and the polarization. When it hits the opposite side of the material it again is divided into a part that reflects back into the material and another that emerges and continues to propagate in the medium. The third thing is internal “excitation” (see Figure 3) and in general is different from what happens with a slab. When the material is large and flat relative to a wavelength (like a slab), this interaction can be modeled as repeated reflections, but when the material is near to or smaller than a wavelength and more compact than a slab, it is called a particle. The internal energy excites the particle like a resonator, which emits like an antenna. Because an antenna is a resonator, the particle can be modeled as an antenna. When the particle is small relative to a wavelength it acts similar to a Hertzian Dipole. When it is larger, things become a bit more complicated.

Regarding particles, the two most important factors controlling the excitation-based interactions are the dielectric differences (between the medium and the particle) and the size of the particle (in comparison to the wavelength of the EM field). Also, there are polarization sensitivities in the scattering process, and the shape of the particle makes a difference.

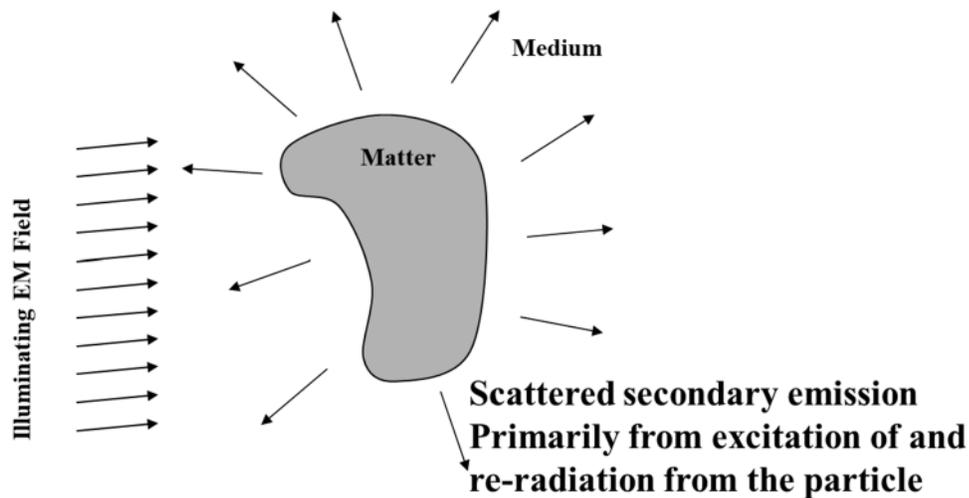


Figure 3. Particles re-radiate energy that enters them

Scattering in any volume is caused by heterogeneity– the “mix” of it. If the EM field encounters a perfectly homogenous medium throughout its path between transmitter and receiver, then there is no scattering. The scale of the mix in a naturally occurring heterogeneous medium can vary widely – consider individual gas atoms being different from the vacuum in which they live, or rain in air, or turbulence on a scale of several meters making invisible boluses of higher and lower pressure air. Regardless of the scale of heterogeneity, the physics of scattering applies.

But, why does matter scatter EM fields? Matter is made up of electrical charges – protons and electrons. When an object encounters an EM field, these charges are set into vibration by the electric field of the EM wave. These vibrations, in turn, radiate electromagnetic waves as a type of secondary radiation, and this radiation we call the excitation mode “scattered wave”. Of course, some of the energy is absorbed by the object, and converted into heat.

This explanation does not work too well for extremely small and high energy particles, such as groups of free protons and electrons. The Ionosphere contains vast regions of such particles, and is quite effective at scattering HF and some VHF signals. That kind of particle scattering is even more complicated than our topic, and isn’t applicable to microwave scattering in the Troposphere by rain.

2 Particle Size vs. Wavelength

One of the significant contributions to theories of scattering has been the work by Lord Rayleigh (pronounced Ray-lee), in the late 1800s. He formulated a basis for scattering from mathematical formulations and observations that show a very strong tendency for scattering to increase dramatically as the wavelength gets shorter, or as the particle gets larger. Mie (pronounced like “me”) extended this work in various papers published in the early 1900s. Mie accounted for the rapid increase of scattering with particle size coming to an abrupt end as the particle circumference approaches the wavelength of the EM radiation. Mie established a ratio χ which is the circumference divided by the wavelength. At $\chi=1$ and larger the particle begins to resonate, which complicates the scattering process in a predictable manner.

It is common to divide χ (scattering to wavelength ratio) into three regions – the Rayleigh Region, extending to wavelengths larger than ten times the circumference of the particle, the Mie or Resonance Region, where wavelengths are between about ten times the circumference of the particle and 1/10 of the particle circumference, and finally the Optical Region from a wavelength of about 1/10 the particle circumference and smaller.

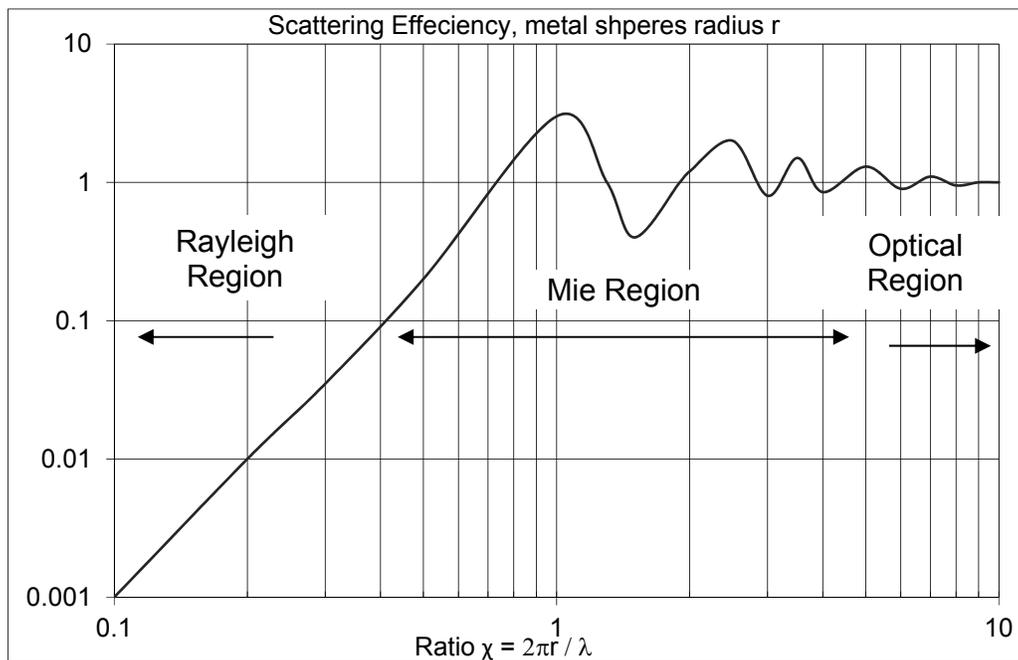


Figure 4. Scattering efficiencies vs. frequency for spheres. Various materials will reach the resonance region at slightly different ratios but the general form of this function is the same regardless of the material.

In the Rayleigh Region, the scattering efficiency increases by the fourth power of frequency. This is a pretty extreme factor, even considering our use of logarithmic scales in communications. The author and others have used rain scatter for communications on 10 GHz, 5 GHz, and 3 GHz, and have observed scattering on 2 GHz, but with very low signal levels. Using the fourth power law - at 5 GHz (about half the 10 GHz frequency) we would expect to receive 1/16 the signal (1/2 to the fourth power) - representing 12 dB less received signal level on 5 GHz than at 10GHz. At 3 GHz (1/3 of the frequency) we would expect to receive 1/81 of the signal, or about 19 dB less signal than at 10 GHz, and at 2304, less than 1/4 the frequency yields 1/256 or 24 dB less signal. These figures are not exactly correct because the dielectric constant for water is changing over these frequencies, and so the scattering efficiency also is changing for that reason as well.

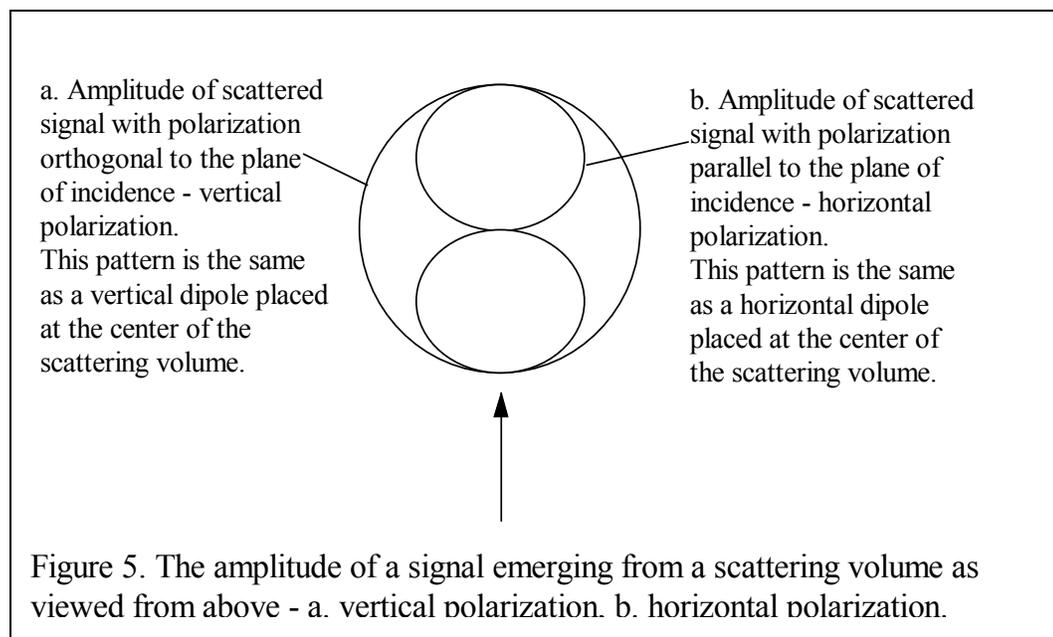
What does this mean for higher frequencies? Unfortunately this geometric increase comes to an end. As the frequency goes up (for the same particle size), we begin to enter the Mie region and the refractive index of water is decreasing. These factors combine to permit rain scattering contacts on most EHF bands under the right circumstances (and when sufficient power is available). Unfortunately, rain is often accompanied with moderate to high humidity (where the dew point approaches the air temperature). As the frequency rises in the EHF part of the spectrum, the volume where there is rain is likely to have considerable absorption due to water vapor. That absorption is likely to reduce the signal level and therefore likely to decrease the chance of using rain scatter effectively.

3 Azimuth Dependency and Polarization

Of some concern to hams wishing to communicate via scattering, especially rain scatter is the directional dependency of the signal. In other words, does the amplitude of the scattered signal drop off at some angles, or is it enhanced at some others? Answers to this question are very difficult in some circumstances and easy in others.

For wavelength/size ratios above about 10, generally speaking, the scattering volume acts like a dipole of the same polarization as the illumination. This means that you can think of the scattering medium as “re-transmitting” the signal as a dipole would if it were centered in the scattering volume. That dipole will have the same polarization as the transmitter. This re-transmission analogy produces correct results for spherical objects (like raindrops) where there is a large wavelength/size ratio (in the Rayleigh Region).

Polarization preservation has consequences. If the transmitter is vertically polarized, then other stations could receive the signal with equal amplitude regardless of their direction relative to the scattering volume. This is the same as receiving a vertically polarized signal from a typically vertically polarized repeater with equal signal levels regardless of where one is located on a circle around the repeater site.



Consider transmitting with horizontal polarization. The “re-transmitter” has the same polarization as the transmitter. If the transmitter were pointing north into the scattering volume, then the scattered signal will look like a dipole oriented east west and radiating mostly north and south. Receiving stations north and south of the scatterer will get strong signals, but stations east or west of it will receive very weak signals or none at all. Stations in other directions around the scattering volume will receive some fraction (as in figure 5). In the last section of this paper, the consequences of the effects of polarization on rain scatter at 10 GHz and higher frequency ham radio bands are presented.

When the wavelength/size ratio becomes smaller than 10, and especially when it goes below unity, the picture changes quite rapidly. Under these circumstances, the forward direction of scattering becomes stronger than back or side scattering. This is true for both polarizations. The polar plots of intensity are different for vertical and horizontal polarizations, but the forward scattering begins to rapidly dominate as the frequency goes up or the particle size goes up beyond a 1 to 1 ratio. Compare some of these plots to 3 or 4 element Yagi’s, or end fire long-wire antennas. In these situations the horizontal plot begins to look more like the vertical one, as you will see below. Note that there are changes in reflection as well as changes in re-radiation. At lower frequencies the dielectric properties of water will cause greater reflection, so the apparent back-scatter will be considerably greater in the 2 to 24 GHz region than in the EHF bands. This backscatter effect is not of much importance to amateurs

who are most often trying to achieve DX (where forward scattering is desired). However, there are circumstances, such as intervening obstacles where backscatter QSOs are desired, and where 2 to 24 GHz communications can be more effective than some of the re-radiation plots suggest.

4 The Atmosphere and Water

Fluids such as gasses (the atmosphere) exhibit some heterogeneity at different scales. Consider the atomic level, where the nitrogen, oxygen and water molecules are essentially disruptions in the vacuum that they are imbedded. Even though they are small compared to a light wave, they do perform some scattering, resulting in color in the sky. At a larger scale, there are dust particles suspended in the air, and there are water particles of a variety of sizes including raindrops.

4.1 Water Particles in the atmosphere

Water in the atmosphere tends to clump into several ranges of particle sizes. Although the possible size of particles completely fills all size ranges from single molecules to the largest raindrop,

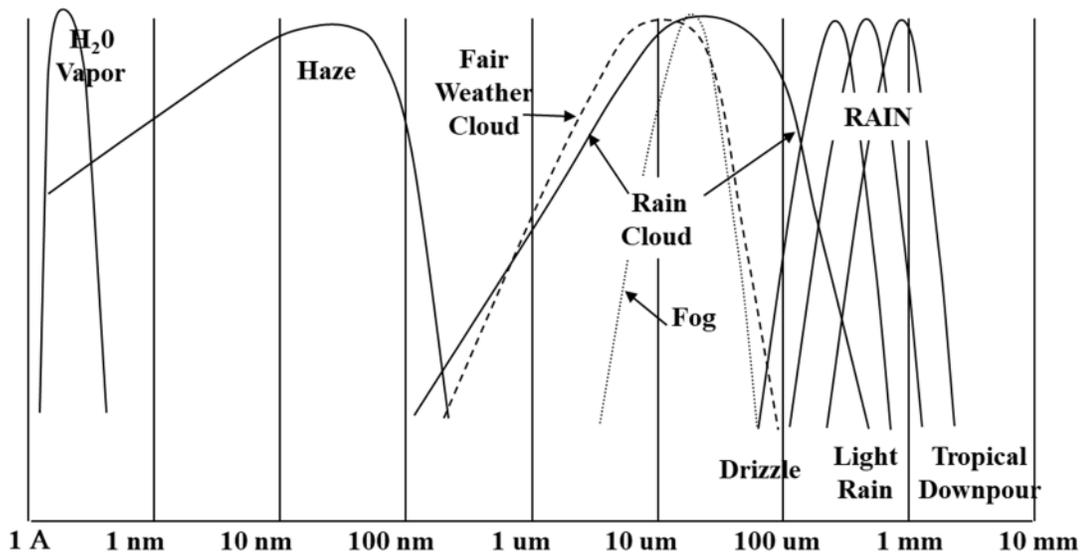


Figure 6. A chart of the distribution of water particle radii for different atmospheric phenomena

our observations of common natural states and our naming conventions describe specific particle size ranges. The first is water vapor - which is individual molecules that have properties similar to other atmospheric gasses. The next larger size is haze particles, which range from under .001 um (a thousandth of a micro-meter, just a few molecules stuck together) to 0.1 um (just a few million molecules), and those in clouds range from .001 mm to 0.1 mm. Fog particles are within the bounds of the size of cloud particles.

Rain droplets range from about 0.1 mm to 1.5 mm (very light drizzle to tropical downpour) radius. Rain is in a good range for scattering of 10 GHz signals. Drop circumferences are 1/50 to 1/3 of a wavelength - and indeed it is the heavier rainstorms (with their larger rain drops) that give the best scattering results.

4.2 The Refractive Index of Water

The refractive index of water is a function of frequency. At low frequencies it is quite high. From 5 GHz to 400GHz it drops from over 8 to about 2. From that frequency to light wavelengths it varies somewhat, eventually falling to a constant 1.33. Also, the refractive index is a function of temperature and changes when the water becomes solid (ice) or gas (vapor). The variety of temperatures found in rain in nature does not significantly affect the figures presented below.

It is probably worth mentioning that refractive indices have both a real and an imaginary part. For simplification, figure 7 shows only the magnitude of the index at a temperature of 20 C. The index changes very slightly at lower temperatures. The plots of scattering presented in Section 5 were all prepared using complex refraction of water at 20 C.

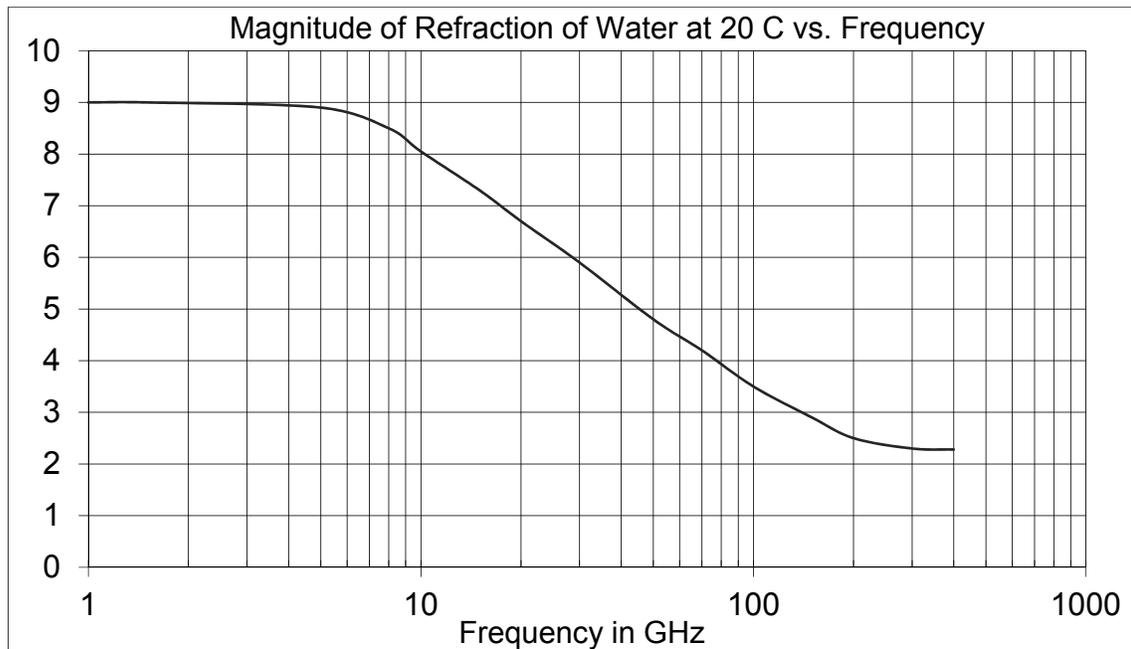


Figure 7. A simplified chart of the refractive index of water changing with frequency.

Tropospheric Scattering

Typical VHF and UHF scattering is caused (for the most part) by turbulence in the atmosphere. This turbulence is manifest in localized pressure differences with sizes that range from a few meters down to about 10 cm. This constitutes heterogeneity at the same scale as the localized pressure difference because different pressures creates different refractive indices. Even though these refractive differences are small, collectively they cause scattering of VHF and UHF signals. Scattering in the Troposphere, which we call “Tropo” scattering from turbulence exists at 10 GHz but it is much weaker than on 1,296 MHz. This paper is mostly about rain scattering. There probably is some contribution to microwave Tropo communications by airborne dust and insects (it has been recorded in Millimeter Wave Radar), but the author has not found a definitive reference on this topic regarding communications.

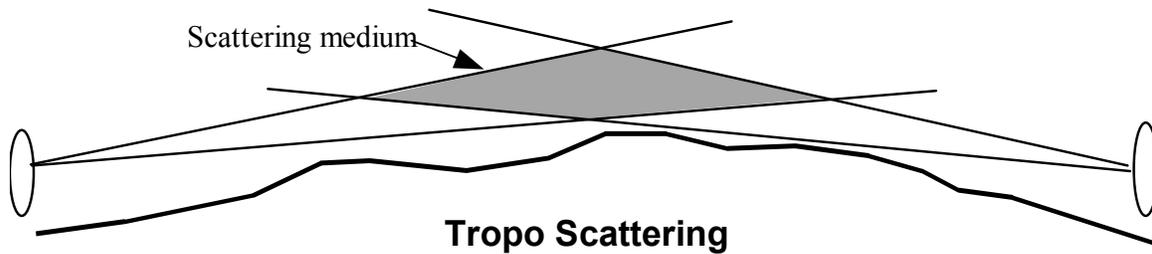


Figure 8. Typical propagation modes for microwave communication

We tend to use fairly narrow beam antennas on SHF and EHF - often with beam-widths of five degrees or less. You can think of the transmitted signal like a light beam from a searchlight. These antennas tend to “illuminate” smaller volumes of the scattering medium than broader beamed VHF arrays. This results in interesting short term fading effects and significant variability in signal levels day to day when the primary propagation mode is Tropo scattering. It is not uncommon to get a 3 to 6 dB boost from a low flying large commercial aircraft that happens to pass through the illuminated volume. The focused beam also allows one to efficiently illuminate a single thunderstorm a hundred miles away, making rain scatter DX a reality.

4.3 Practical use of Rain Scattering

On narrow-band microwave operations, we normally use horizontal polarization, which is parallel to the plane of incidence. Scattering for this condition follows the pattern shown above in figure 5. This means that in normal amateur radio microwave communications during rain scattering conditions, with everything else equal, some storms are in better positions than others (see Figure 9). The poorest scattering signals are apparent when the rain event is at a point that causes the two stations to aim at 90 degrees relative to one another. At shallower and greater angles, the signals increase. If amateurs want to maximize the likelihood rain scattering, then either vertical polarization should be used or rain cells that do not cause this 90-degree relation should be sought.

On a number of occasions the author and other New England stations W1FKF and W1RIL (now SK) have measured rain scattered signals of significant amplitude and over surprising distances. In the case of signals between W1RIL and W1FKF, where there is a fairly difficult obstruction (a nearby hill), rain scatter has offered over 10 dB and sometimes as much as 20 dB of signal boost. This can easily make the difference between not hearing the other station at all and being able to copy clearly.

One characteristic of rain scatter is Doppler shift. This phenomenon is manifest in a shift in frequency when the path between the transmitter and receiver is shortening or lengthening. Because the rain in a storm is being blown by winds, the scatterer is moving, the path length is changing, and the frequency is shifted. Usually, a CW note becomes quite “fuzzy” in its sound because the individual raindrops are moving at slightly different speeds and directions. The sound is very much like Aurora propagation on 6 or 2 meters - at times it can sound like strong noise. When looking for a rain scatter contact it is advisable to have the transmitting station send CW dashes, so that the receiving end can distinguish the noisy signal from the background noise.

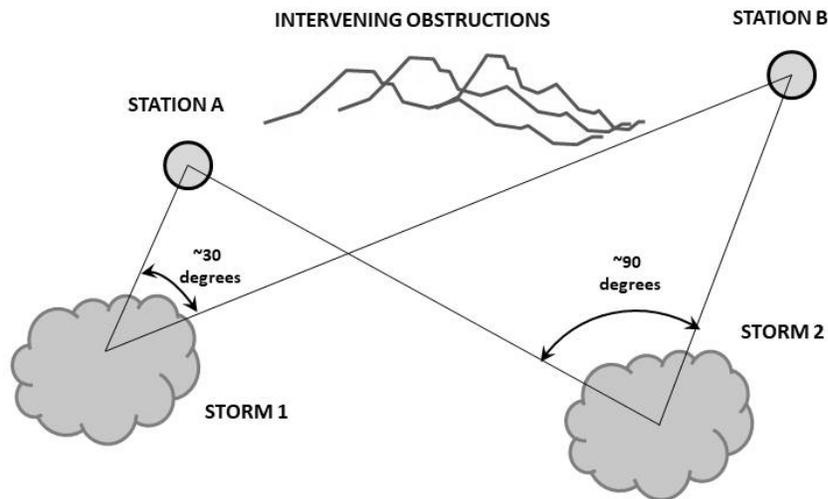


Figure 9. An example Situation. The two stations are much more likely to communicate via storm 1 than storm 2 because storm 2 presents a 90 degree angle.

Because the wind will have one predominate direction, stations will notice some Doppler shift, even though the signal sounds quite fuzzy. Sometimes the wind speed is high enough to move the signal out of the pass-band!

In May of 1995, W1RIL and W1FKF pointed towards a developing thunderstorm with a round-trip distance of some 250 miles and noticed a Doppler shift of 4.1 KHz - which represents a wind/rain velocity of over 130 mph. As it turned out, this storm developed a devastating tornado. On another occasion, the author and these stations scattered their signals off a storm with a round trip distance in excess of 350 miles, and with signals strong enough for a phone QSO. Other stations have reported similar extended DX from rain scattering.

Over the years, we have come up with various ways of trying out rain scatter. If you are lucky enough to have a sked soon after your local TV station weather report, just look at the radar map - it will be the most current and local of any sources. You are more likely to need an up-to-date map at other times of day, so consider obtaining a radar application or weather application that includes weather maps for your cell phone. There are some programs designed to help amateurs use rain scatter (see references). Remember, when the two stations' paths to the storm are at 90 degrees, the signals will be weakest, but try anyway.

4.4 Snow Scattering

Snow scattering is very much like rain scatter, and at 10 GHz has approximately the same scattering effect for the same equivalent rainfall rate. However, most snowfall occurs at a slower equivalent rainfall rate than typical rain, so only the heavier snowstorms are likely to have significant snow scattering. Nonetheless, the author and other users of 10 GHz have observed snow scattering. At times it can be very effective at extending communications even though it is unlikely that hilltop expeditions would take advantage of this mode. Some of these sites are cold even in the summer!

Many rain events, especially in the cooler months and in more gentle circumstances than thunderstorms create an effect called "brightening". The precipitation starts at high altitudes as snow. As the snow descends through warmer air it begins to liquefy slightly and the snowflakes clump together to make much larger wet flakes. This only happens in fairly calm atmosphere. Eventually these might re-freeze or melt to rain as they fall further towards the ground. At the layer where the large clumps are formed the particles are quite large and form a mirror-like surface. Although these tend to be fairly low in the atmosphere, they can provide very strong scattering like no other

phenomenon. They also exhibit little or no Doppler shift as they only occur in relatively calm conditions. Perhaps brightening layers are not going to set records for scattering DX, but they can provide excellent signals.

4.5 Rain Scatter on 10 GHz and above

Although the author has participated in amateur radio contacts on EHF, none have used rain scatter. Rain scatter has been demonstrated numerous times with specialized cloud radars, which operate at 35 and 95 GHz. Such radars can distinguish between different size water particles in the cloud, and can easily characterize raindrops, ice crystals and sleet. This is all done with backscatter, even though forward scatter can be stronger than backscatter at some frequency/raindrop size ratios.

The plots below show directional intensity for the upper ham bands at both polarizations and at different rainstorm conditions. In all cases the transmitter is at the left, and the plot is a view from above the scattering volume. The plots show both the vertical and horizontal polarization. The color version available in the file might be easier to interpret than the printed version here.

You might notice when carefully comparing plots that the same wavelength/size ratio does not always produce the same pattern. This is because the refractive indices of water change with frequency. Also, with a wide range of 40 or so dB in the chart it is possible to contain all the plots for a given frequency except that for the 10 GHz plot, the smallest raindrop size plot detail is lost. Therefore, a separate chart for 10 GHz and 0.5mm drops is provided.

The three raindrop sizes chosen correspond to the average rain drop size found in the three storm intensities with narrow distributions around those sizes. Plots show these three drop size effects separately on the same chart mostly to provide the reader with a handle on the effect of drop size at frequency. However, actual rain storms carry a distribution of drop sizes. A storm with narrow distribution around 0.5 mm raindrop radius is at the mean size of “drizzle”. This kind of storm would have rainfall of under 0.25 mm per hour (under 1/100 inch per hour). The 0.9 mm drop radius mean size would indicate a moderate storm of about 12.5 mm per hour (about ½ inch per hour), and the 1.4 mm drop radius characterizes the mean size of only the very heaviest of thunderstorms and tropical downpours of 100 mm per hour (4 inches per hour). A further plot is provided where size distribution covers the entire range representing the typical distribution of a moderate rain rate of 12.5 mm/h with a range of drop radii from under 0.2 to over 2 mm. This separate plot is meant to characterize a typical moderate rain storm. Very light rain storms and very heavy ones would be different.

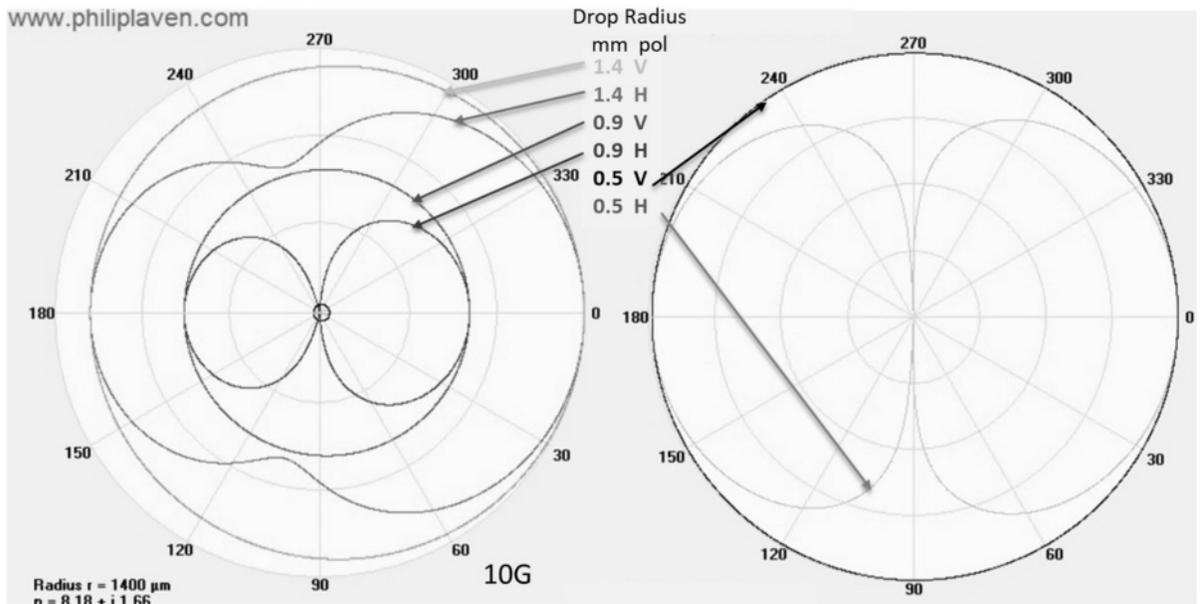


Figure 10. All these plots assume that the incoming EM wave is from the left. On 10 GHz it was impossible to see the shape from the smallest drop size on the same plot as the larger, but on all the others they “fit”. In this figure the smallest drop size plot is a small circle on the left plot, but is expanded to an entire plot by itself on the right.

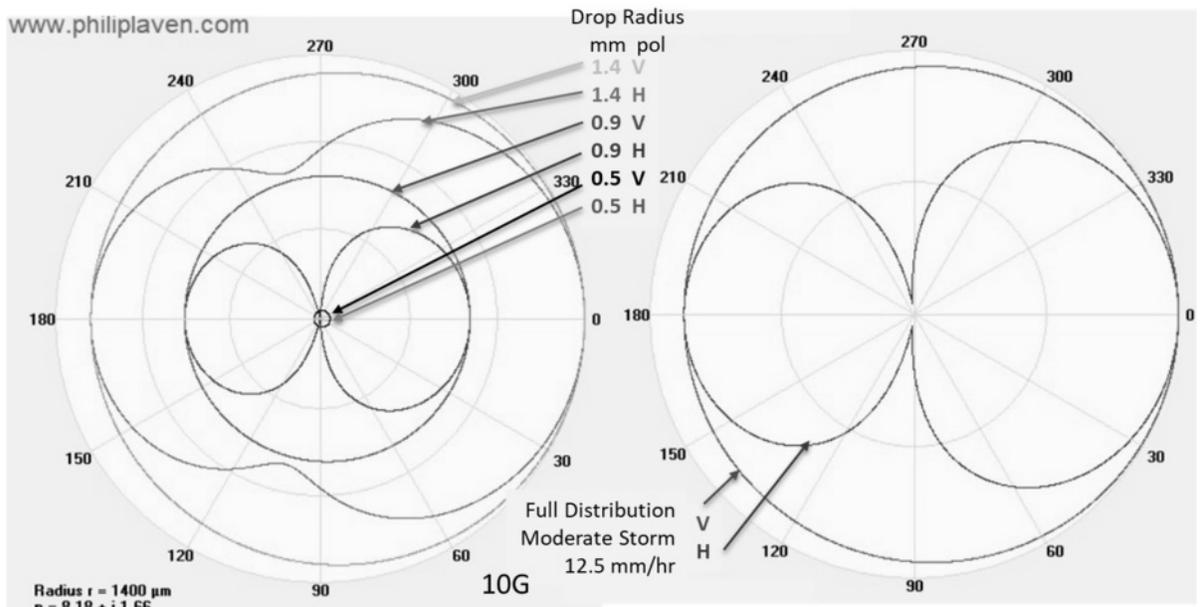


Figure 11. On this plot and all below, the left plot shows the scattering diagram for three very narrow ranges of raindrop size which are the mean of drizzle, a moderate storm and a tropical downpour. This helps visualize the differences caused by rain drop size. On the right is a more realistic plot that represents scattering from a typical distribution of drops from a moderate storm that rains at a 12.5 mm/hr rate (1/2 inch per hour).

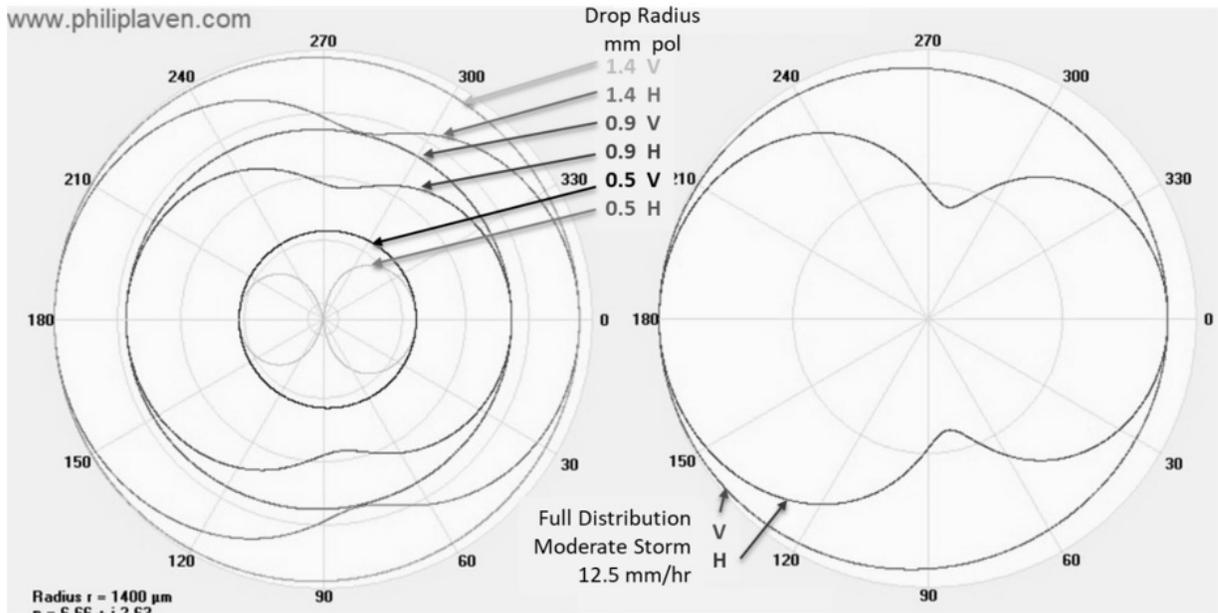


Figure 12. 24GHz plots of scattering. On left, three drop sizes in two polarizations. On the right, scattering with a drop size distribution as found in a moderate rain storm of 12.5mm/hr (1/2 inch/hr).

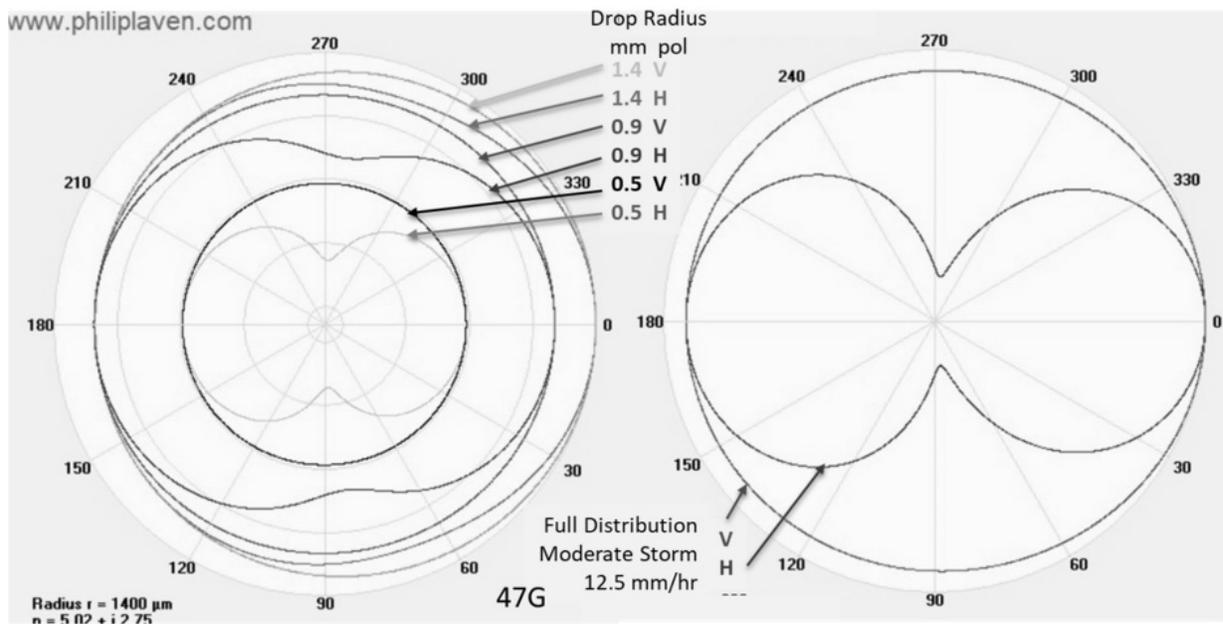


Figure 13. 47GHz plots of scattering. On left, three drop sizes in two polarizations. On the right, scattering with a drop size distribution as found in a moderate rain storm of 12.5mm/hr (1/2 inch/hr).

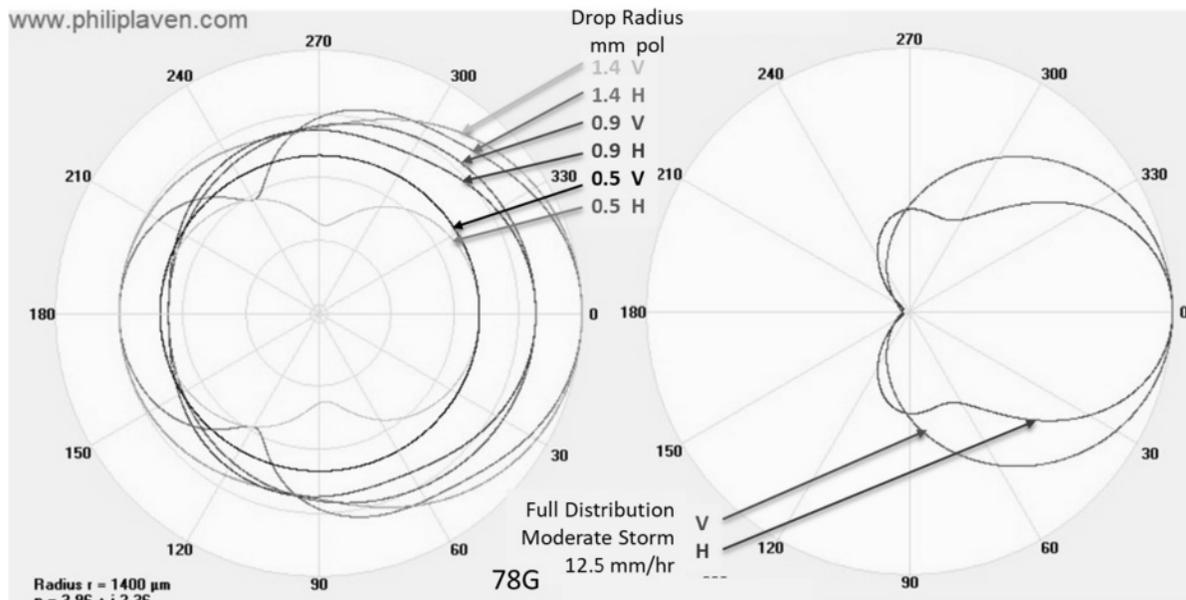


Figure 14. 78GHz plots of scattering. On left, three drop sizes in two polarizations. On the right, scattering with a drop size distribution as found in a moderate rain storm of 12.5mm/hr (1/2 inch/hr).

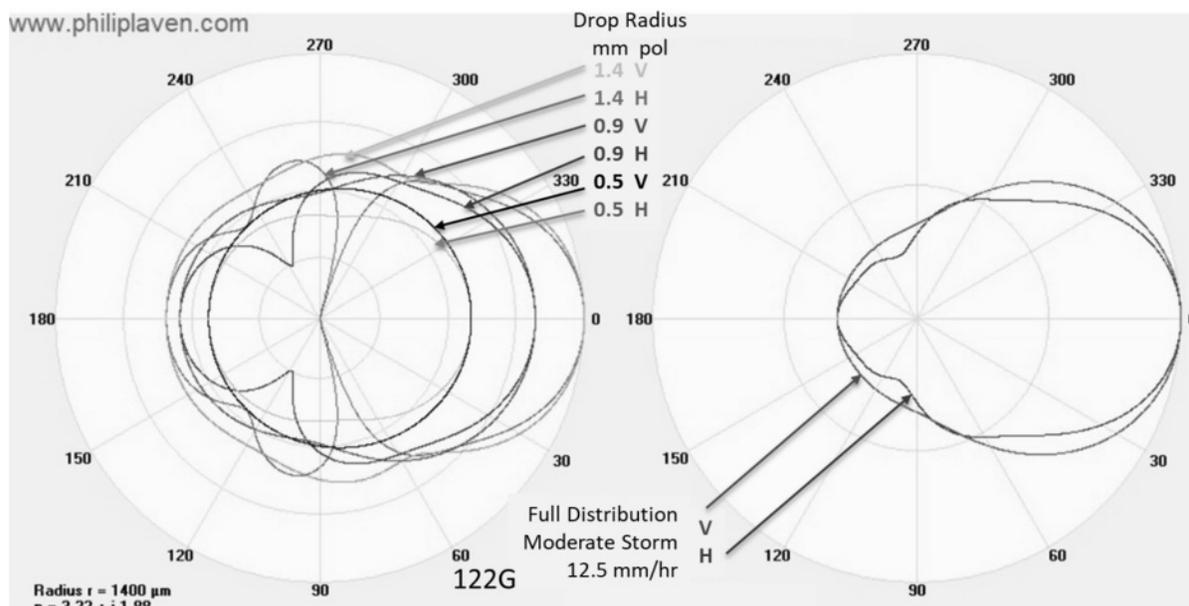


Figure 15. 122GHz plots of scattering. On left, three drop sizes in two polarizations. On the right, scattering with a drop size distribution as found in a moderate rain storm of 12.5mm/hr (1/2 inch/hr).

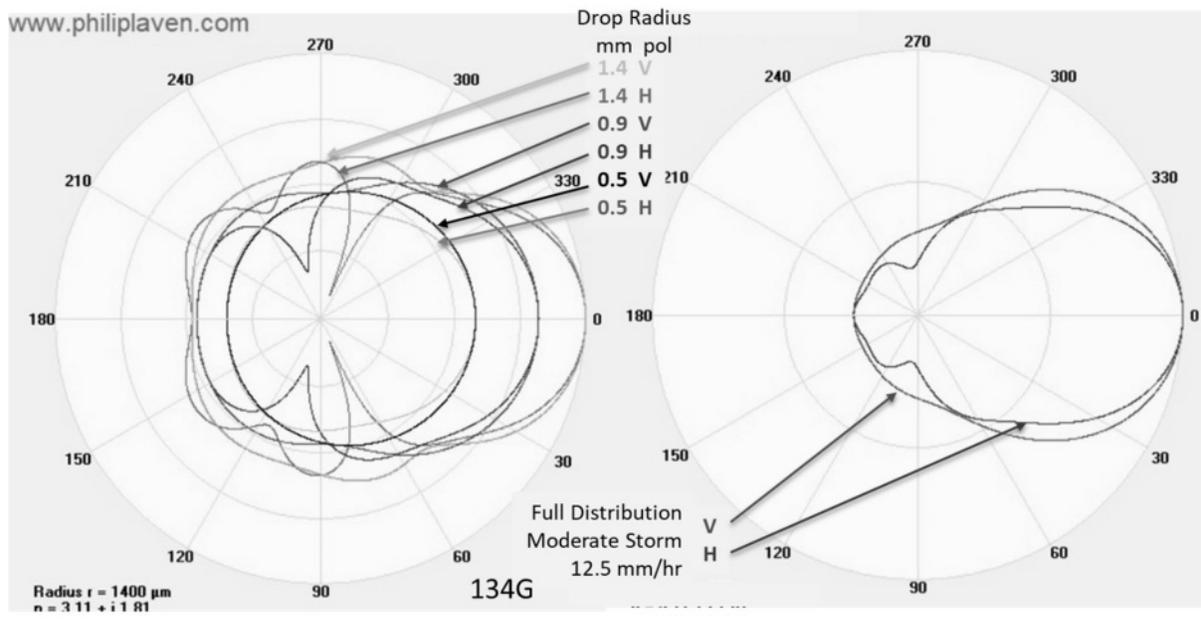


Figure 16. 134GHz plots of scattering. On left, three drop sizes in two polarizations. On the right, scattering with a drop size distribution as found in a moderate rain storm of 12.5mm/hr (1/2 inch/hr).

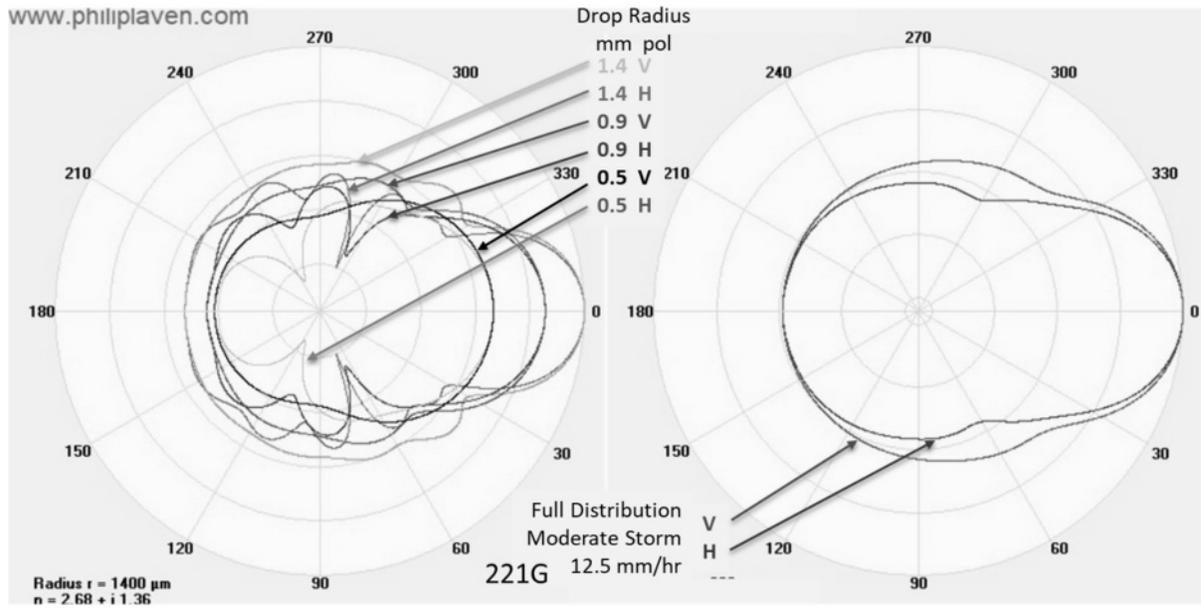


Figure 17. 241GHz plots of scattering. On left, three drop sizes in two polarizations. On the right, scattering with a drop size distribution as found in a moderate rain storm of 12.5mm/hr (1/2 inch/hr).

5 Attenuation

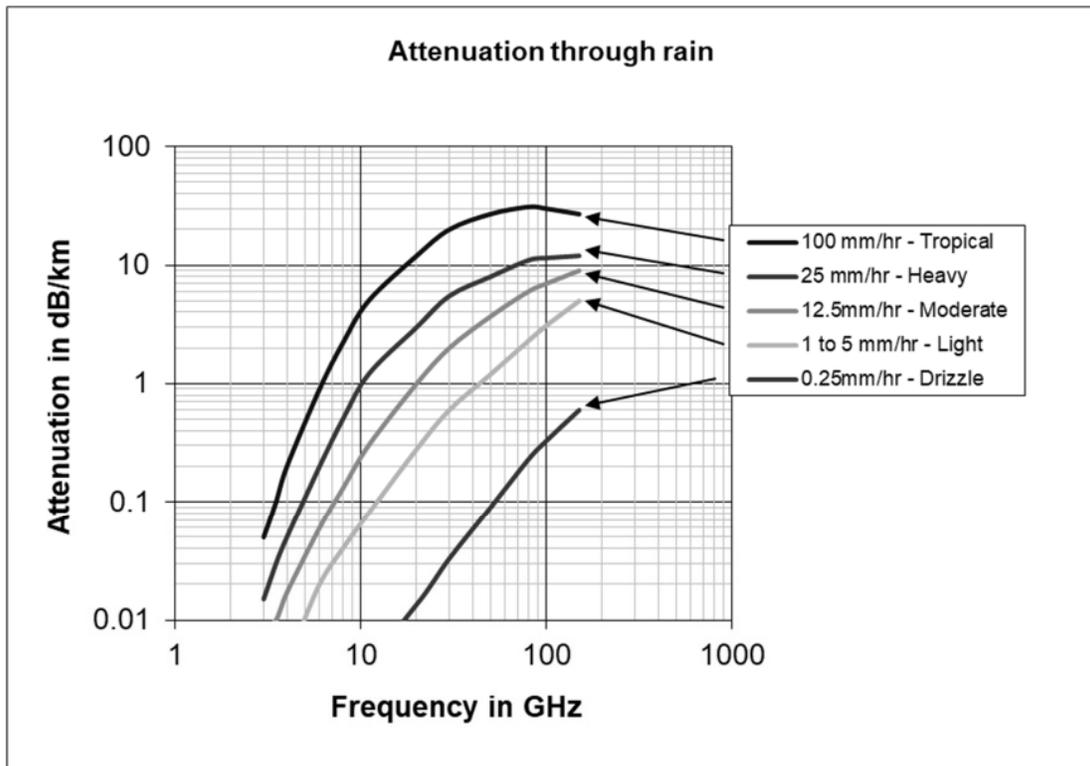


Figure 18. Attenuation through rain at different rain rates, according to total water content and frequency. This does not take into account atmospheric loss. It is due to rain alone.

Unfortunately, scattering alone does not tell us how well the signals will be heard at the other end of a link. Significant attenuation can occur in rain, especially as the frequency rises. Figure 18 includes information that the author could acquire, and does not include attenuation levels below about 0.01 dB/km, or frequencies below 3 GHz and above 145 GHz. These data should be useful in determining attenuation in most SHF and EHF situations.

As can be seen from Figure 18, significant path loss can occur through rain, especially heavy rain, which is the very condition that will improve forward scattering at EHF. Fortunately, thunderstorms are usually quite small, only covering a few kilometers. Therefore, the best situation for a rain scatter enhanced EHF contact would be a small and intense thunderstorm rain cell located midway between the two stations. Table 1, below, gives calculated data for scattering through a moderate storm where some atmospheric losses are included. This atmosphere was calculated at 50% relative humidity at just under 60F. Typical storms in the summer may have such characteristics, but to be fair, many will be at a lower temperature and higher relative humidity. The important dew point is likely to be around the range of this example, so the table is likely to be useful in practice as a first approximation.

Table 1. A table of effective attenuation in db/km of signals through atmosphere with rain falling at various rates. Here the atmosphere is set to 50% relative humidity, and the air temperature to 288 Kelvin, 15 Celsius, 59 Fahrenheit. The dew point is 5 Celsius, 41 Fahrenheit. Actual summer storm conditions may have a similar or higher dew point and lower atmospheric temperatures.

dB/km	Rain Rate mm/hr													
GHz	0.5	1	2	4	5	10	15	20	25	30	40	50	75	100
10	0.019	0.027	0.045	0.088	0.112	0.248	0.403	0.571	0.749	0.937	1.336	1.760	2.908	4.157
24	0.214	0.287	0.435	0.732	0.881	1.629	2.379	3.131	3.884	4.639	6.150	7.665	11.459	15.263
47	0.556	0.816	1.276	2.090	2.468	4.199	5.777	7.263	8.685	10.056	12.687	15.202	21.142	26.736
78	0.996	1.444	2.181	3.391	3.926	6.255	8.258	10.077	11.769	13.366	16.351	19.128	25.462	31.209
122	1.757	2.308	3.180	4.560	5.157	7.690	9.810	11.700	13.435	15.056	18.047	20.793	26.950	32.439
134	1.654	2.217	3.104	4.501	5.104	7.653	9.778	11.668	13.401	15.018	17.996	20.726	26.835	32.267
241	3.308	3.892	4.798	6.204	6.805	9.317	11.388	13.215	14.880	16.427	19.262	21.846	27.586	32.652

6 Conclusion

So, what does this information about propagation mean to a ham? One thing is that a home station can be lots of fun, especially if there are other hams with home stations within 30 to 100 miles. It also means that portable operations can work in situations that would normally seem impossible - such as where there are mountains in the way. Rain may be a nuisance to operating, but it can be a welcome signal enhancer that lets you work around mountains and more than a hundred miles beyond normal range. Although moisture is usually the biggest enemy at EHF, one small intense thunderstorm, or a calm region of rain that provides the brightening effect, fortunately placed will probably enhance communications, and possibly be a propagation mode for record setting DX.

A very similar paper was presented at Super VHF Conference, in April 2019, Sterling Virginia.

FOR MORE INFORMATION

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Mie, G., (1908). Beitrage zur Optik truber Median speziell kolloidaler Metallosungen, *Ann. Phys.*, 25, 377-445

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Rayleigh, Lord (J.W. Strutt)(1871), On the light from the sky, its polarization and colour, *Philos. Mag.*, 41(107),274.

Ulaby, Fawwaz T., Richard K. Moore, and Adrian K. Fung, Microwave Remote Sensing Active and Passive, Volume 1 (See chapters 4, 5, and 6), Artech House, Norwood, MA, 1981. All three volumes of this fine work are highly recommended for serious millimeter wave interests.

From Andy Flowers, K0SM:

<http://www.n5dux.com/ham/files/pdf/Using Radar Data to Predict Rain-Scatter Paths.pdf>

and an application devoted to Amateur Radio Rainscatter Activity

<http://www.frontiernet.net/~aflowers/rainscatter/>

Lightwave Communication: Why now is the best time to try it

By Warren Ferber – WF0T & Rob Jahnke – K0XL

How it all started

In 2016, I presented a new idea for Lightwave Communication at the Northern Lights Radio Society's Aurora conference. Up to this point, the state of the art was high-power LED's attached to unusually complex modulation schemes, using Fresnel lenses mounted in heavy square wooden enclosures, and mountain tops. These systems work really well, see Clint Turner's (KA7OEI) site modulatedlight.org and you'll read about his work in setting a U.S. distance record of 273Km using visible red light.

I wanted to try something different, I wanted to experiment with IR light. I found a 5mm IR LED at Digikey that had a built in 3-degree half beam width lens for under 60 cents. I decided to start with this LED as the basis for my testing. Modulation would use a small microcontroller called an ATTiny85, prepackaged on a circuit board from Digispark. These can be programmed using the Arduino IDE. I wrote a simple program to send my callsign and grid square via CW then a series of tones that I used to test the audio response of my receivers.

I built an 850nm (352 THz) 4 LED beacon. I ran various tests including straight distance and cloud bounce with it and had positive results.



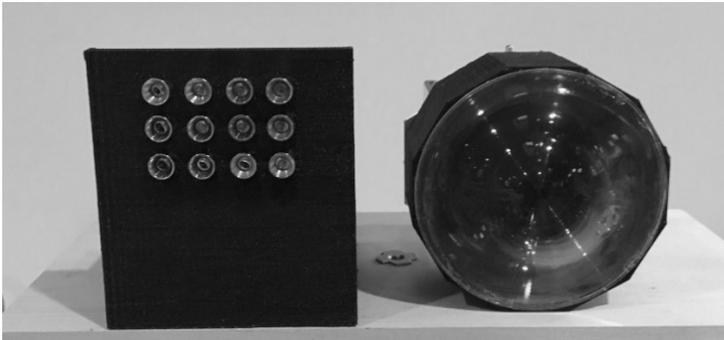
4 LED IR Beacon

- SFH4550 IR LED's
- Digispark ATTiny85 microcontroller board
- Modulated at 50% duty cycle
- Callsign, Grid locator, and tones from 23hz to 4.6Khz
- Radio Shack plastic project box
- 9volt supply

With a 5-inch glass lens and Clint's v3.10 optical receiver I successfully received my reflection off clouds up to 7000 feet and overdrove my receiver from 1.5Km in direct testing. It was fun and exciting, and I wanted to take this experiment to the next level and create a working Transmitter and Receiver.

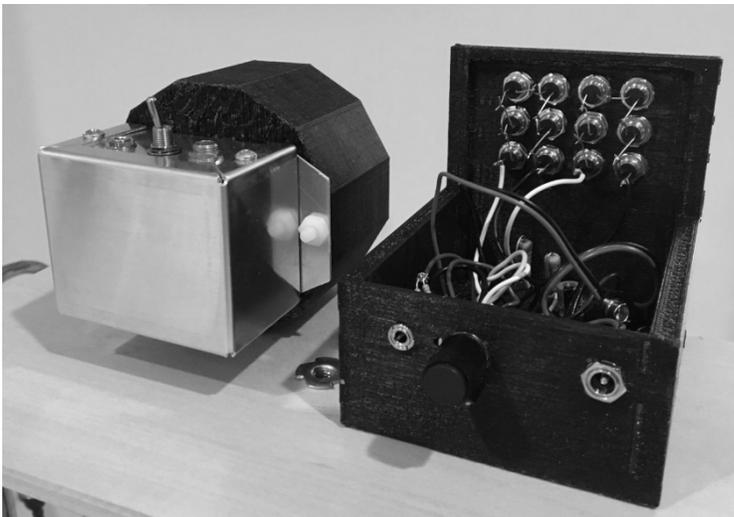
I purchased a 3D printer “kit” based on a review in Maker magazine, called the Cetus3D mk2. This kit came mostly built. I screwed down the print plate, attached the cables connecting the 3 motors and print head, and attached the vertical slider for the Z axis. I decided the best way to learn 3D printing was to jump in the deep end. I designed a lens enclosure first. When that design failed, I decided maybe wading in the pool was a better idea. I designed and printed some simple shapes, combined a few until I got the settings where things looked good, then went back to designing a lens enclosure. That was successful, so I went on to the transmitter box.

The results can be seen below.



Transmitter and Receiver

- 555 timer PWM modulator
- 50% duty cycle
- LM833 based audio amplifier
- 12 – SFH4550 IR LEDs wired in 2 parallel blocks of 6 LEDs in series
- K7RJ circuit board of KA7OEI’s v3.10 optical receiver
- 3D printed lens holder and transmitter box
- 90mm Fresnel lens with a 50.8mm focal point

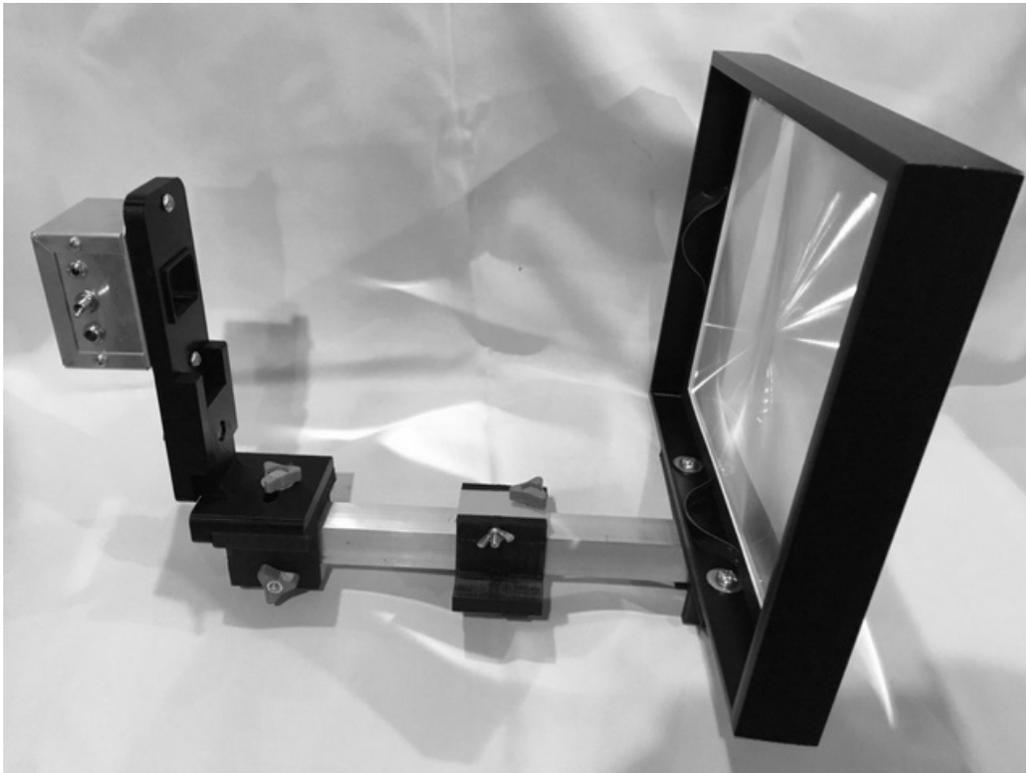


This was a repeatable build. I printed a couple of these and used them for multiple tests. It produced a significantly brighter (in IR light) rectangle that allowed me to; receive music reflected off clouds at over 9000 feet, complete my first voice lightwave contact with Donn WA2VOI, and a water tower bounce test. This was an inexpensive build, \$2.20 in PLA filament and 3 hours of print time. Unfortunately, it was not a stable platform for long range communication, but was good enough for distances found in Minnesota. This is an easy to build and aim system for getting your feet wet in lightwave communication.

Back to the drawing board

It was about a year of work to get to the current design. After exhausting several different designs, I was starting to think that large wooden boxes may be the way to go for a stable platform. I was cleaning up my workspace when I ran across my large format rail camera.

This was a stable photography platform that has many adjustments including focusing the lens with the film board. Maybe this is what I was looking for, so I got to work designing an Optical Rail Receiver. The results are below. This is actually version 2 as it includes some of the improvements that Rob K0XL made. They are noted below.



From left to right; Optical receiver, K0XL 3-axis receiver mount, tripod mount with rail tensioner, and the fresnel lens mount, all mounted on a 1-inch square aluminum boom

There have been several iterations of the various parts, the rest of this white paper will cover the build process to create your own Optical Rail System for transmitting and receiving.

After settling on a set of parts, assembling them, and waiting for dark it was time to test it out.

Success! The platform was stable at multiple angles up to 85 degrees, the maximum my tripod can go. There was no change in the focus between the lens and receiver. I was always able to keep my test LED with its tiny spot centered on the optical sensor at all times. Testing continued throughout the evening, including bouncing IR light off various trees, garages, and street signs up and down the blocks near my home. In every test the light from my optical transmitter

reflected onto the optical sensor as expected. It was time to share this design with my optical colleague Rob Jahnke, K0XL to get his opinion of this new design.

Rob saw it and agreed that this was the stable platform that we needed for serious optical light work around the area. He liked it so much that he made some modifications and improvements that are incorporated in my build.



Rear view showing K0XL designs

This view shows the optical rail tensioner and modified receiver mount designed by Rob. This view also shows how the receiver box is mounted to the rear of the vertical slider.

Rear view; Optical Rail Receiver showing mounting of Hammond Manufacturing box #1411FBBU



WF0T 3-axis receiver mount

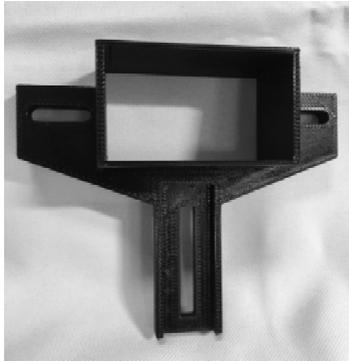
I wanted to build-in the ability to set the appropriate focal length for the lens while retaining finer adjustments to the vertical and horizontal axis. This would allow us to quickly and easily insure that the incoming light through the fresnel lens could be centered on the optical sensor.

During testing in the field, it was surprisingly easy to make focus adjustments. During the water tower testing, the red beacon on the top of the water tower made a fine red dot that could be minimized to the smallest size and centered on the sensor. This also could be peaked for maximum desensing of the audio output from the receiver.

The mount is designed for boxes with mounting between 3 1/4" to 4 3/4". This is Version 3.1 and includes a light shield to block sunlight and nearby street lights.

Front view of WF0T v3.1, 3-axis mount with receiver

Here are the individual parts for WF0T v3.1, 3-axis mount.



Vertical slider w/light shield Horizontal slider Tripod base mount Fully assembled



Front view of WF0T v1, 3-axis mount with transmitter

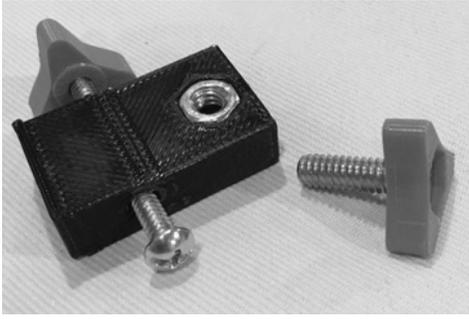
WF0T Transmitter option

The v1 3-axis mount can be used for the transmitter and LED. The Hammond box used for the receiver is perfectly sized to hold the microphone amplifier and my new ATtiny25 modulator for the transmitter or the older 555 timer based modulator.

Fine tuning adjustments can be made to focus the LED beam using an appropriate Fresnel lens mounted to the lens mount.

This allows for various tests you will want to do because it is fun! Cloud bounce, water tower bounce, snow bounce, etc.

Both versions of the 3-axis mount use $\frac{1}{4}$ inch hardware for mounting the sliders to each other as well as to the tripod base mount. A 1-inch hex bolt will press-fit into the tripod base mount. Tightening the horizontal slider can be secured with a simple wingnut.



K0XL Optical Rail Tensioner

One part that has proved valuable was one that Rob created due to a problem he was having with my original rail mounts. The lens mount was tippy. He solved it by creating these little tensioners that slip in the mounts.

It was a month later that I found out why he was having this problem. He was printing with 30% fill, while I was printing with 65%. My mounts were just solid enough to hold. As soon as I started printing with 20% fill (my printer's setting) I saw the same problem Rob noticed. I now use these as well.

K0XL 3-axis Receiver Mount

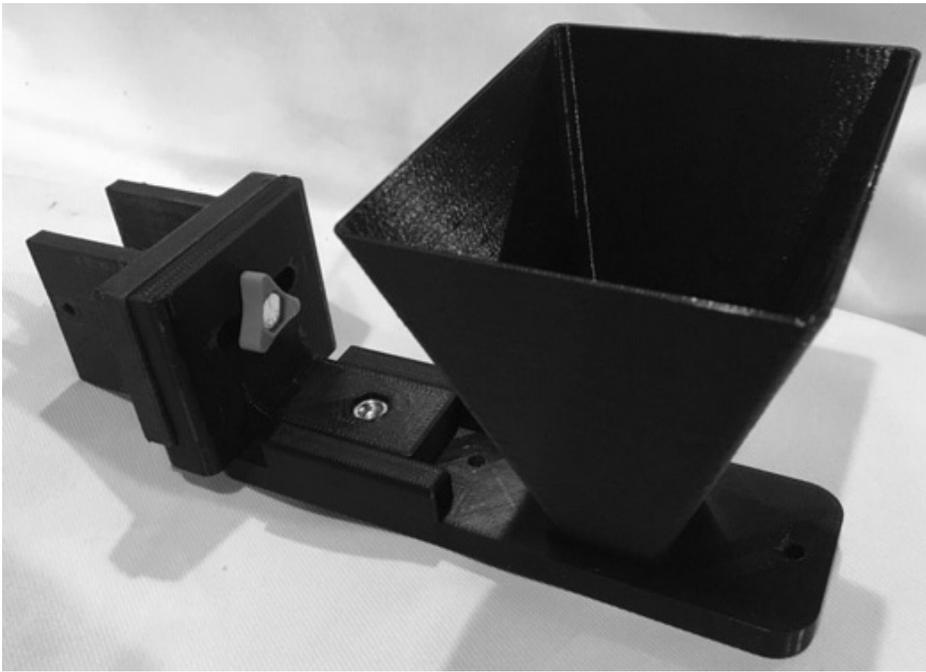
This is Rob's version of the receiver mount. His design was based on wanting a way to shield the receiver sensor from outside light sources. The added benefit is that this form appeals to the NLRs membership due to its 10 GHz-like form.



K0XL version of 3-axis receiver mount and light shield

The parts fit together using ¼ inch hex bolts. You can use wingnuts to secure the sliders together and for mounting to the base mount, or you can use the knobs that Rob designed.

The light shield needs to be glued to the receiver opening. The shield was made for a "page-sized" Fresnel lens. These are available at CVS, craft stores, Ebay, etc. Most of these sized lenses have a focal length around 13 inches and the shield is optimized for this.



Fully assembled K0XL 3-axis receiver mount with attached light shield



Individual parts for K0XL 3-axis mount; vertical slider, horizontal slider, base mount

WF0T Lens mount

I wanted to create a lens mount that would be able to support any size lens I had. Over the last few years I have accumulated many different sizes and shapes. What I came up with was a mount that supports a picture frame. The frame can be any size needed to match your lens, within reason. I have tested it with 12-inch x 12-inch frames, down to 5-inch x 7-inch frames with positive results. The lens mount remains solid on the rail with no noticeable wiggle. I have used it outside in winter wind conditions and the test signal remained focused on the center of the sensor.

I used frame kits that I purchased from Michael's crafts and Dick Blick Art supply. The are metal frames that come with two sides, so you can customize your frame to your lens. I then cut a photo mat to size and use that to hold the lens in place. The frame kits come with metal clips.



Bottom view of Lens mount attachment

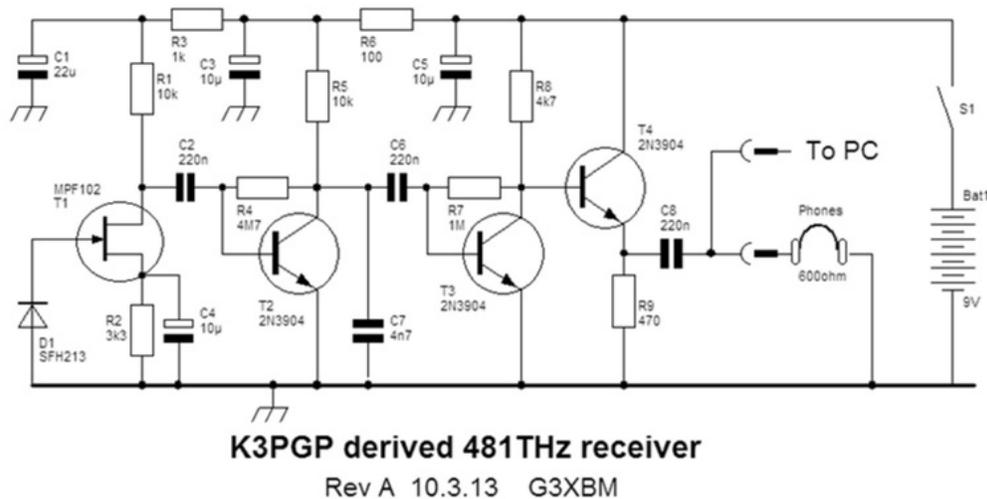


I use ¼ inch hardware to secure the frame to the mount, bolts, washers, and lock nuts. It creates a solid platform for holding the lens as well as making it easy to change the lens size.

Receivers for lightwave communication

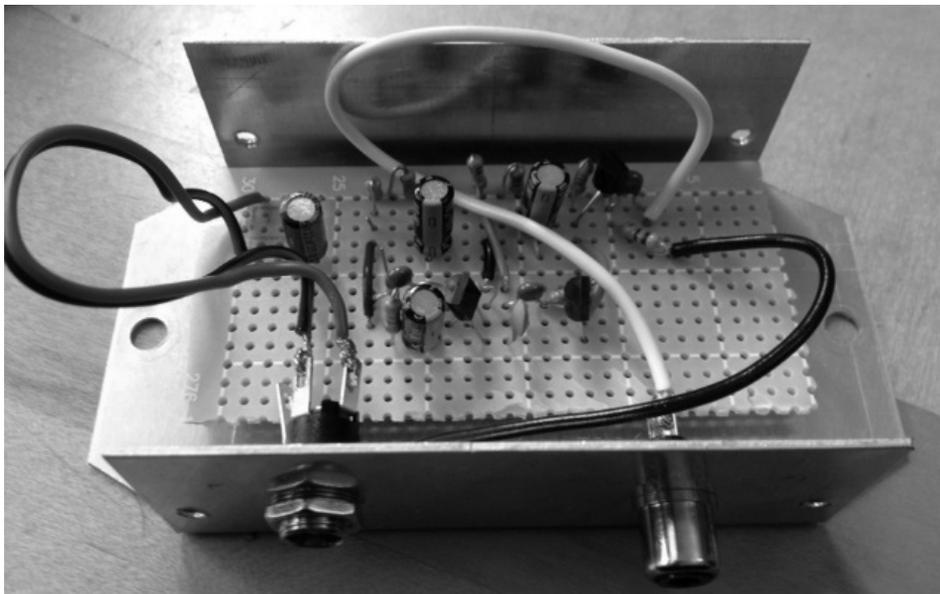
There is a lot of information on building optical receivers on the internet, everything from a solar cell connected to an audio amplifier to serious designs for long distance communication. I wanted to include a couple of options that we have tried and have had success with.

The first is a design from G3XBM, it is derived from the original K3PGP optical receiver. I used this schematic to build from which I got from the VE7SL blog.

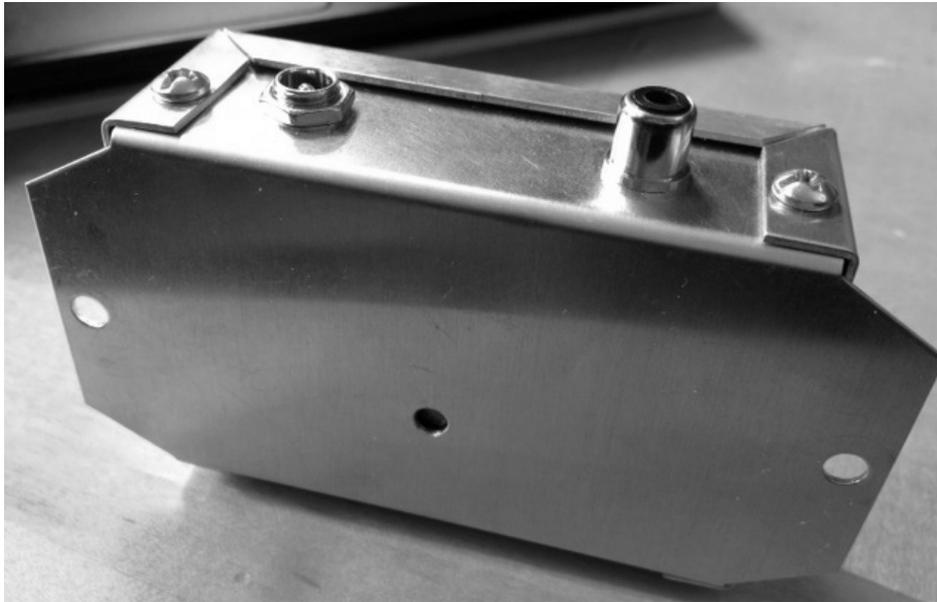


I made a couple of parts modifications. I used a BPW34 detector instead of the SFH213, I also replaced the MPF102 JFET with a 2N5457, and used a 1 meg resistor on the base of T2 instead the 4.7 meg.

I built the receiver on perf board. All but two of the parts are visible on the top of the board. I mounted the JFET and detector on the bottom of the board facing the front of the enclosure.

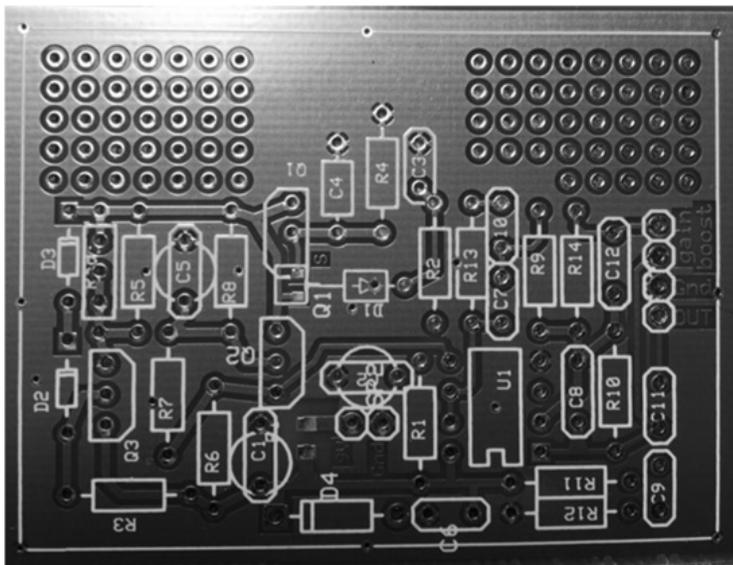


This is the finished receiver in the enclosure. I built it in a Hammond Manufacturing aluminum box #1411FBCU. The opening is for the BPW34 detector. It is connected “in the air” to the gate of the JFET. This minimizes the capacitance and increases the sensitivity of the detector. On top are the connectors for power (left) and audio-out (right).



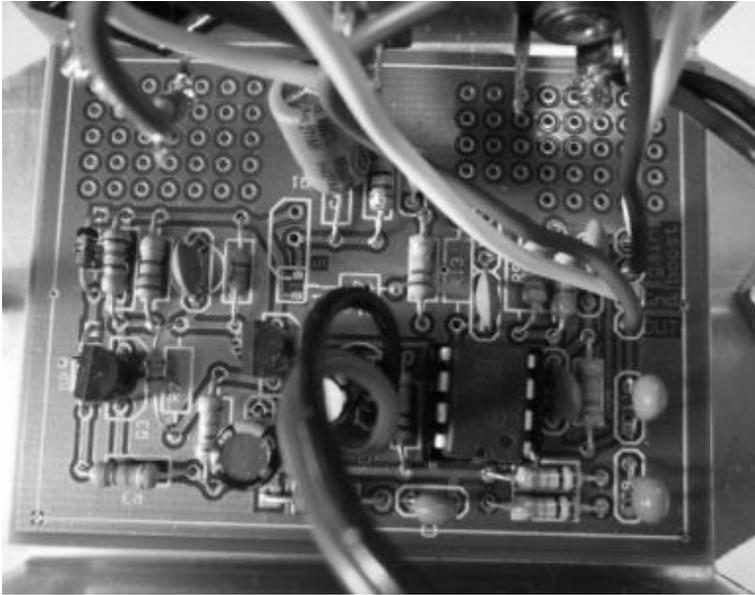
I have received aircraft strobes over the horizon and visible aircraft above 25,000 feet and over 50 miles away. I also did my first cloud bounce tests with this receiver and was able to detect my beacon off of a cloud deck at around 7,000 feet. It is simple to assemble, goes together quickly, and really works. G3XBM designed a super optical receiver!

The second receiver we are using is one designed by Clint (KA7OEI) the v3.10 optical receiver. Available on Clint’s site (modulatedlight.org) is the version that Ron, K7RJ, created a circuit board for. The PCexpress files are available on the site. When Rob saw this he put in an order for 6 boards for us to build and to use in our testing.



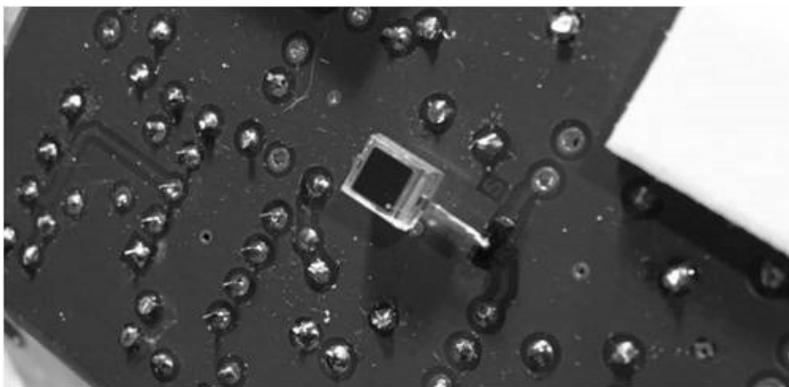
Clint's site has many pages of build hints, so will steer you there for specifics about the design and optimizing your build. I will share some observations on my builds and results of our testing.

There are only a couple of surface mount parts on the board. One is a fuse to protect against applying power to the wrong terminals in the dark, and the JFET. The rest of the parts are standard sizes. The board allows for a pretty quick build, I completed my second board in under two hours, mostly due to trouble getting the FET and detector mounted horizontally to the board. Mounting it into the Hammond box with drilling and aligning took another hour.



Populated circuit board

Mounting the Detector to the JFET can seem daunting; however it isn't as hard as it may seem at the beginning. A magnifying glass, tweezers, good lighting, and decaf were essential for my success in mounting them on the board. I probably spent more time than I needed on it, but I wanted to make sure the JFET was vertical and the detector was horizontal to the board.



Close up view of the smd JFET and the BPW34 detector, the blob upper right is the double-sided mounting tape

The circuit board and wiring fit in the Hammond box like they were made for each other. Even if you can't find the box I used, the WF0T version 3.1, 3-axis mount, will accommodate the box you do use.



Receiver box, from left to right: 9V battery plug, gain boost switch, and audio output

Results of testing the receivers

Both Rob and I built two KA7OEI v3.10 receivers. Rob adhered to Clint's suggestions on building the receiver. I was not so disciplined; I used whatever parts I had on hand, mostly picked up at a local surplus shop. The amazing thing was all of our receivers using the BPW-34 detector were within 1.5 dB of relative output of each other. The best was the one receiver built using the BPW-34S detector which has a built in daylight filter. The G3XBM receiver was also tested and had the highest audio output.

WF0T KA7OEI w/daylight filter	0 dB -57dB	+20 dB gain switch -35dB
WF0T KA7OEI no filter	0 dB -55dB	+20 dB gain switch -37dB
WF0T G3XBM	0 dB -29.5dB	
K0XL KA7OEI receiver #1	0 dB -55.6dB	+20 dB gain switch -37.3dB
K0XL KA7OEI receiver #2	0 dB -54dB	+20 dB gain switch -37dB

Table of receiver relative output testing results

K0XL “Flashlight” Optical Transmitter

This is based on the Atomic Beam CREE “white” LED 5000 lux flashlight.

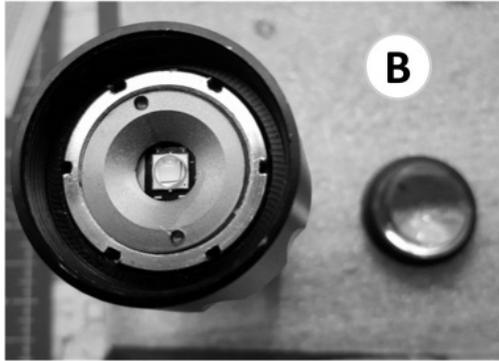


Taking the Atomic Flashlight apart

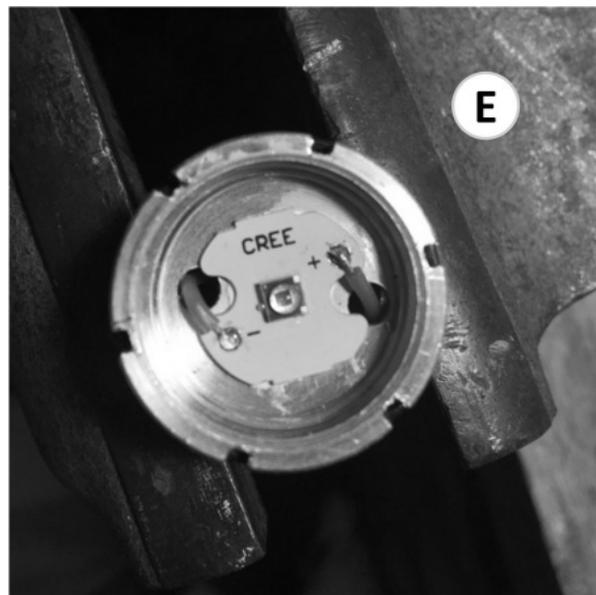
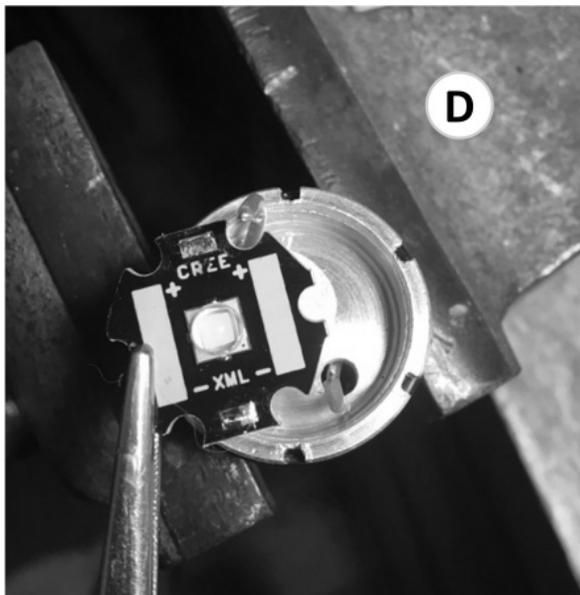
You will need a spanner wrench (or snap ring pliers) to disassemble the flashlight. Start by removing the backend battery cover and removing the battery holder. (A) The switch is press fit into the backend, remove it by pushing on the silicon button.



Unscrew the lens from the lens holder assembly (B). Using the spanner wrench remove the LED assembly (outer aluminum ring) from the flashlight through the lens holder assembly (C).

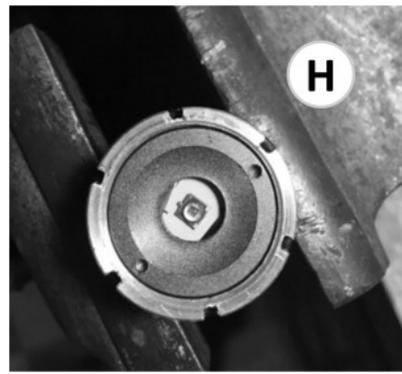
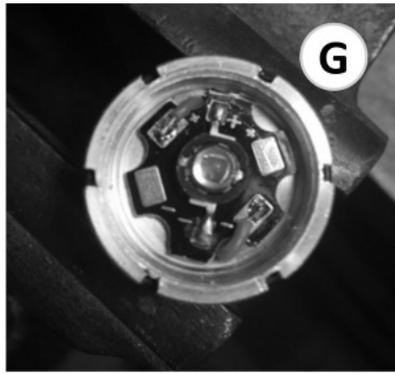
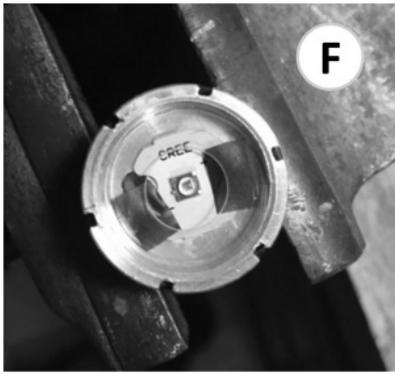


Remove the copper colored (inner ring) LED retaining ring. There is a clear spacer under the retaining ring, keep this as it protects the LED connections from shorting out. The LED and heat sink are now exposed. Unsolder the red (+) and blue (-) wires from the LED base (D). Clean off the LED assembly base and add new heat sink compound and add your new IR or red LED and solder it in place (E).

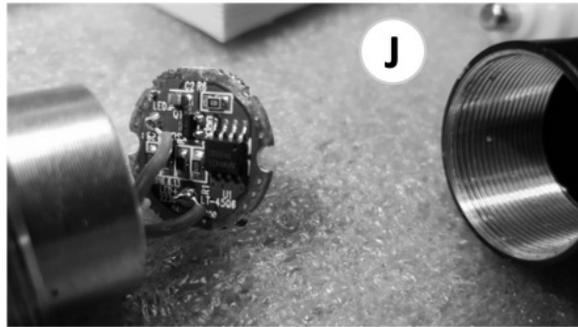


A word of caution is needed at this point. Depending on the LED you are using and the base it is mounted on, you may need to make a small modification to prevent shorting out the LED.

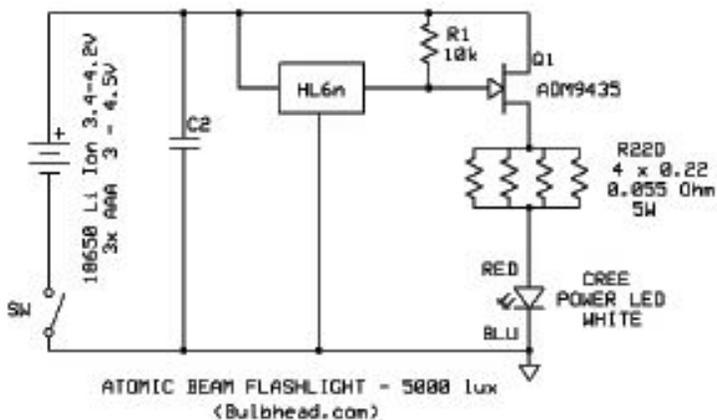
In the first example, my IR LED is mounted on a 16mm Cree base. The positive and negative solder points are slightly exposed under the clear spacer opening. I used small pieces of electrical tape to protect from shorting (F). In the second example, the LED is on an elevated mount with a lens (G). In this case you will need to drill out the copper colored ring to 12mm. This should give you enough clearance to not short out the LED. When you have the LED centered and are happy with the clearance, screw the copper colored LED retaining ring into the LED assembly (H).



Next, let's modify the LED control board. Turn the LED assembly over. Using a tweezer or one pins on the spanner wrench, insert it into one of the openings between the circuit board and the LED assembly and pry to circuit board out (I).

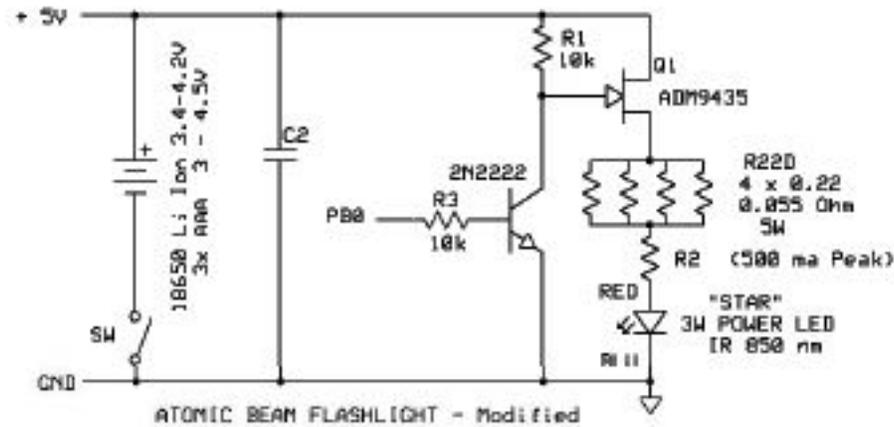


This is the LED control board (J). Part numbers may be different from the schematic; we found part ids were different between shipments. Thankfully it isn't a complicated schematic and the board is easily modified.

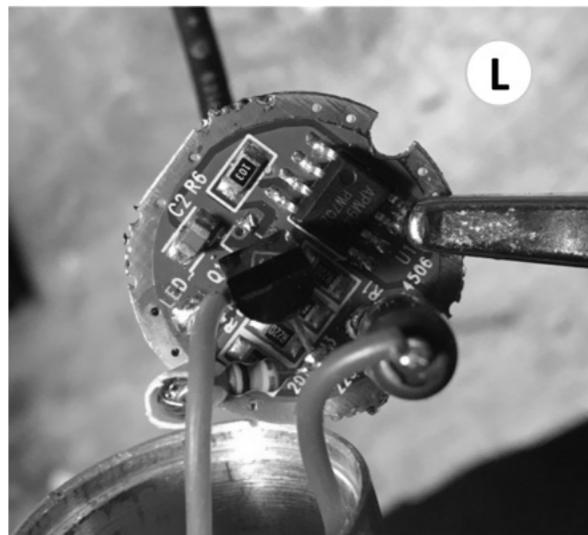
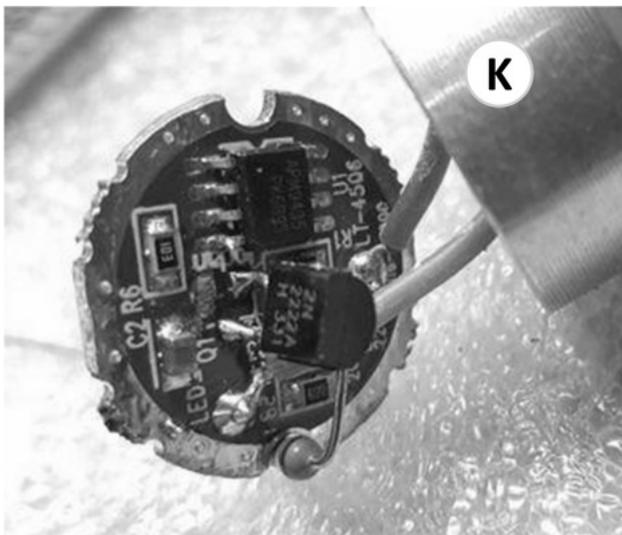


Schematic for the board in the Atomic Beam Flashlight

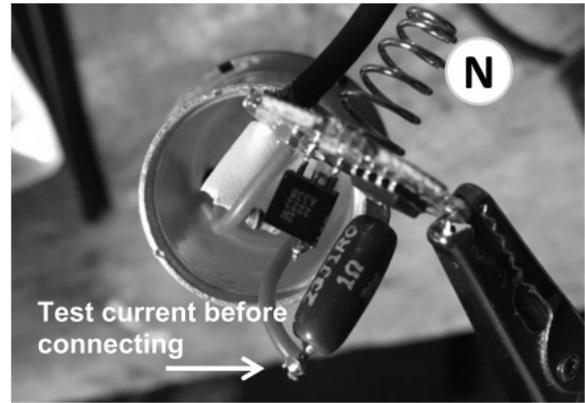
This is the schematic of the modified board. You'll notice that the 3 pin transistor (HL6N) is removed and replaced with a 2N2222A transistor and a current limiting resistor (R2) is added.



You'll need to remove the 3 pin smd device. After removing it, put a dab of solder on the 2 pads we'll use for 2N2222A transistor's collector pin (left pad) and emitter pin (lower right pad) (K), bend the base pin up to the back of the transistor (L).



The transistor will be used to feed the PWM output to the LED. Connect a 10k resistor to the base pin with the other end routed to the back of the board (K). I attached the resistor horizontally to the base pin (L) and routed a length of RG-174/u to the resistor through the semi-circular opening on the side of the board. Shielded cable is not necessary, however it doesn't hurt either and I figured some shielding on the feed line couldn't hurt. The ground is attached to the ground ring (outer ring) on the backside of the board (M).



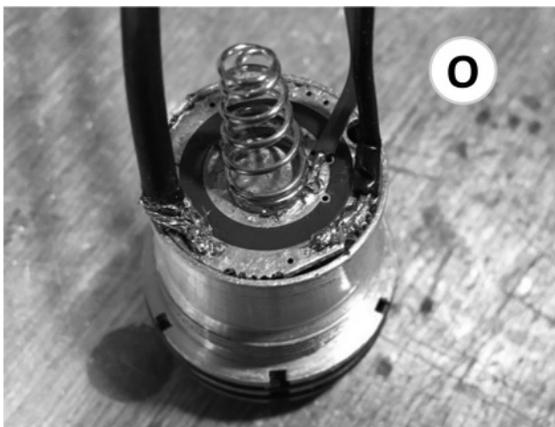
Remove the red (+) wire from the board and attach a current limiting resistor to the positive board pad and reattach the red (+) wire to the other end of the current limiting resistor (N).

Calculating the value of the current limiting resistor is an Ohms law exercise; $R = V/I$. Since we are using a 50% duty cycle a little explanation is needed. First you will need to know, or at least be able to estimate, the 100% duty cycle or DC current limit of your LED. Then you will need to know the voltage drop across the LED.

The new equation is $R = (V - V_{LED}) / (2 * I_{LED})$. For the Cree 3watt IR LED it came to a resistor value of 1.4 ohms. I have a batch of 1 ohm, 3watt resistors and found one with a 1.2 ohm resistance and used that.

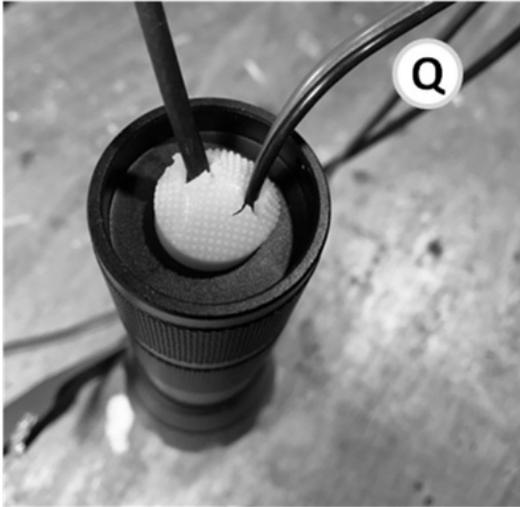
Use the power calculation to make sure your resistor can handle the watts that need to be dissipated by the resistor. Calculate that using $P = (V - V_{LED})^2 / R$. My power calculated out to 2.75 watts, since we are using a 50% duty cycle, I really only need half that amount (~2watt resistor), so my 3watt resistor will be fine.

Attach your power cable to the backside of the board (O). The inner ring is for V+ and the outer ring is ground. Before fitting the board back into the LED assembly test that everything is working and verify the current draw. Connect the output of your modulator to the feed line attached to the 2N2222A. I checked current by putting my meter between the current limiting resistor and the positive connection to the LED (N).



When you are happy that everything is working as expected it's time to reassemble your new lightwave transmitter. Start by press fitting the board back into the LED assembly (O).

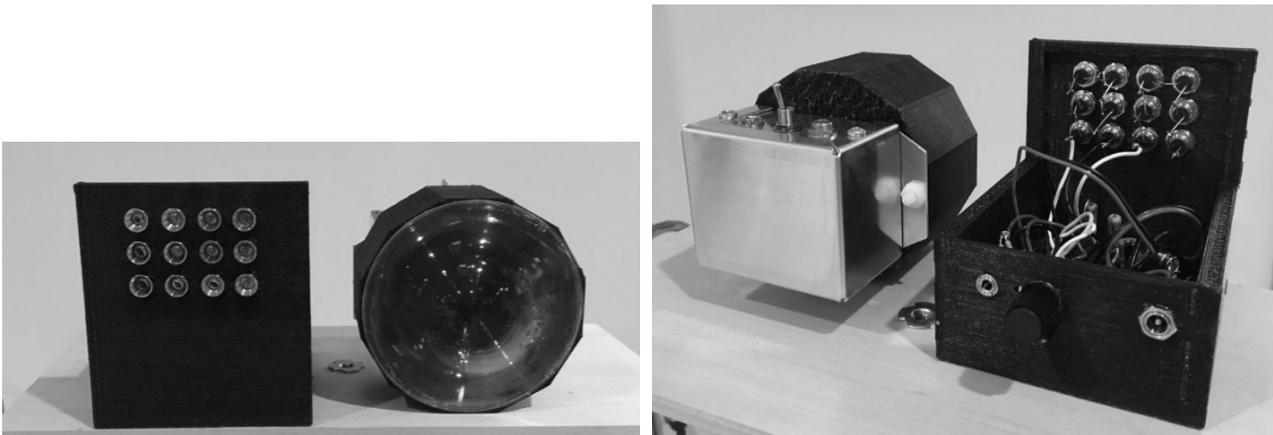
Next, slide the lens holder assembly over the body of the flashlight and using your spanner wrench screw in the LED assembly into the flashlight body. I tied a loose knot in the feed line and power wires to use as a strain relief (P). I then screwed on the backend battery holder. I cut a couple of slits in the silicon button cover and pushed the wire through them, then pushed the button cover into the backend battery holder (Q).



Congratulations! Your new lightwave transmitter is ready.

WF0T 3D Printed Transmitter and Receiver

I introduced this set at the beginning of the paper. Here are a few more details about building the transmitter. The receiver is the KA7OEI v3.1 attached to a 3D printed lens housing.

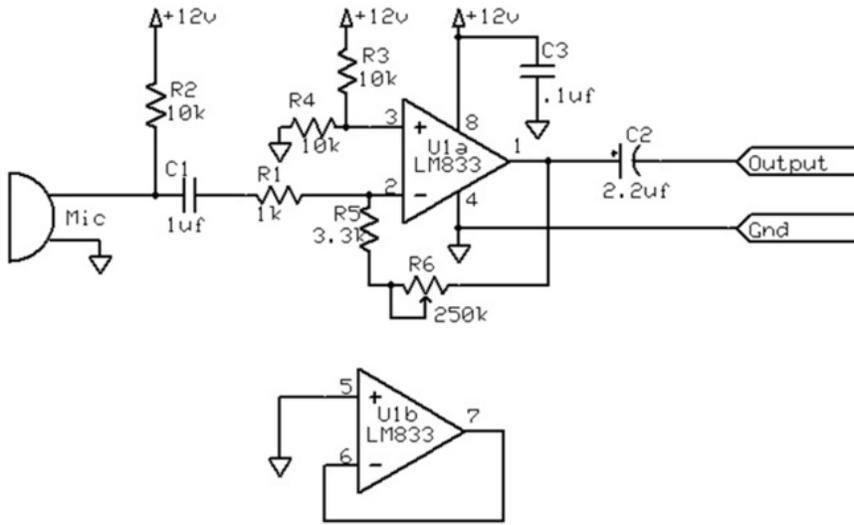


I have uploaded my transmitter and receiver designs to Thingiverse.com.

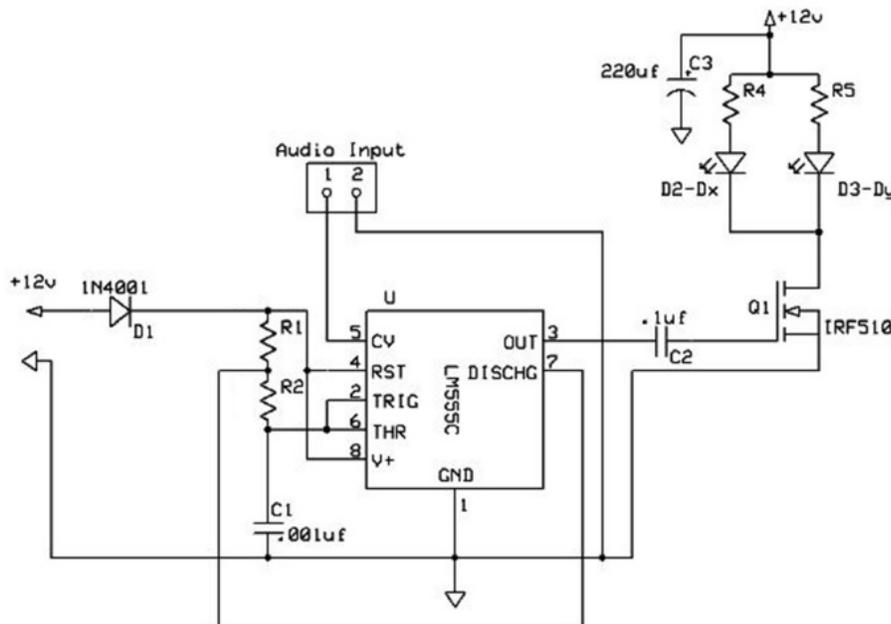
The Transmitter is at: <http://www.thingiverse.com/thing:2751923>

The Receiver is at: <http://www.thingiverse.com/thing:2765972>

I used an LM833 OpAmp in my transmitter, mostly for two reasons; I had a bunch on hand, and I liked the sound it produced. You can substitute any dual OpAmp (TL082, LF353, TL072, etc.) for the LM833 in this schematic and get good results. This is the schematic I drew up for the microphone amplifier.



This is the schematic for the 555 timer modulator. R1 and R2 should be selected to make the duty cycle as close to 50% as possible. I used values of 1.2k (R1) and 56k (R2), they got me really close to 50%.



The 3D printed transmitter uses 2 parallel banks of 6 LEDs in series. I've denoted this on the schematic as D1-Dx for the first 6 and D2-Dy for the second 6 LEDs. R4 and R5 are valued to limit the current to the LEDs based on the voltage drop of the 6 LEDs and the duty cycle of the timer. I'm using 2 - 4 ohm 3 watt resistors. Keep C3 connections as short and direct as possible. If you want to add more bypass for the circuit, you could also add a bypass capacitor after the 1N4001 (mis-labeled as D1) directly to ground.



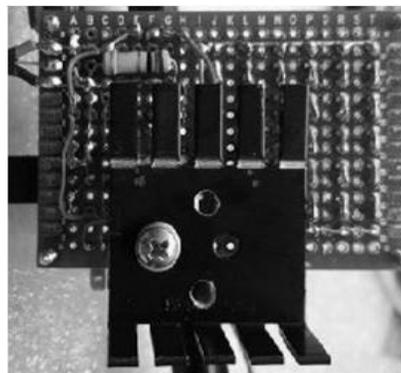
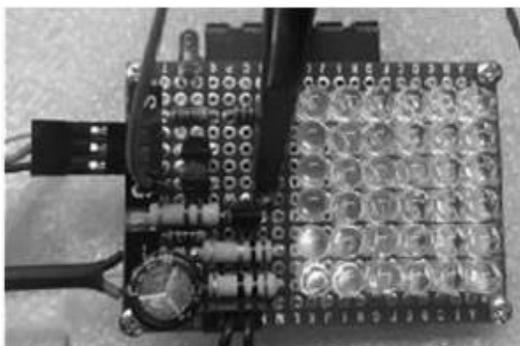
As you can see, it is not the prettiest of builds. Hidden under the wires are three small circuit boards, one for the microphone amplifier, one for the 555 timer modulator, and one for the LEDs driver module. It may be a bit ugly, but it works!

The LED holders are from Radio Shack. Fortunately, they are available on-line from various retailers. They add a finished look and provide some heat sinking for the LEDs.

The LEDs are Osram model SFH4550, they have a 6 degree beam width, which creates a tight pattern of IR light with no external lens required.

K0XL 36 LED Transmitter

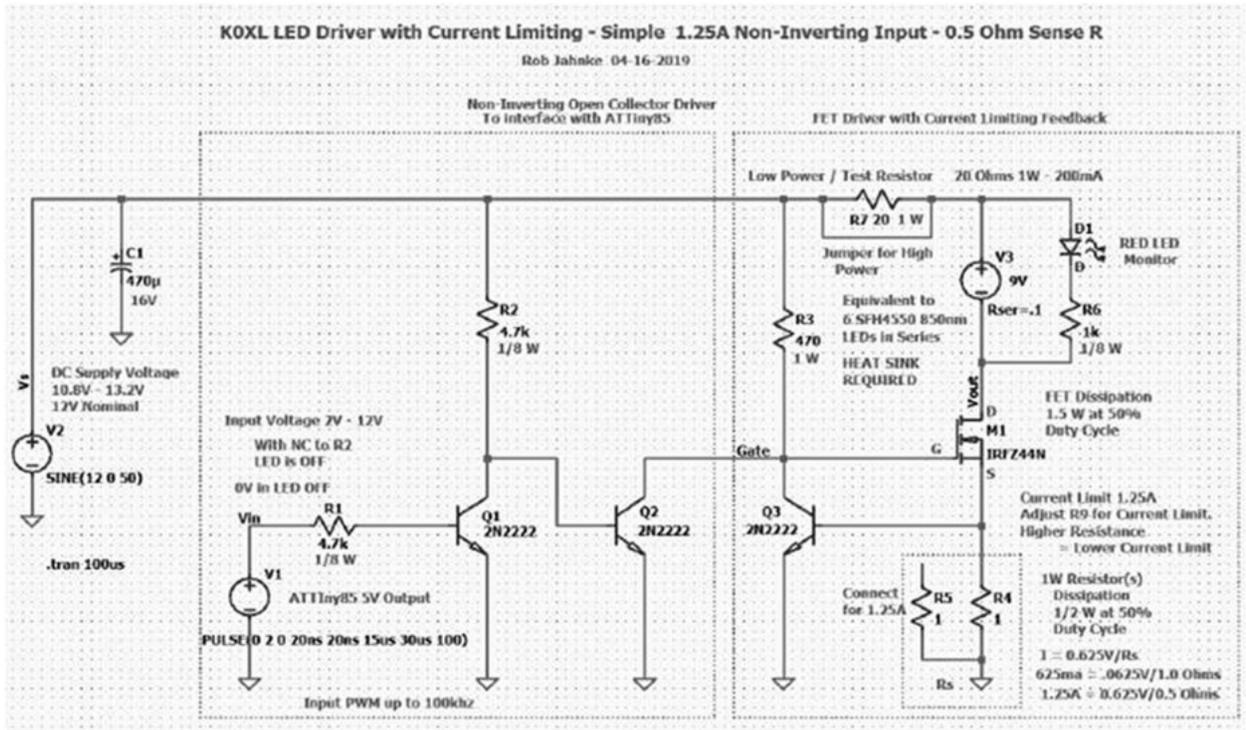
Rob extended the idea of a 12 LED transmitter to 36 LEDs. He built it on a perf board and tightly packed the LEDs into a 6x6 square.



Front side of 36 LED Transmitter with current driver, back of board with FET heat sink view

The LEDs are wired in 6 parallel blocks of 6 LEDs in series. Current to each block is 200ma, driven by Rob's PWM modulator. Heat is the main issue to address with this transmitter. With the LEDs tightly packed airflow needs to be provided. A small PC processor fan will work.

Rob designed a constant current source to power the transmitter. It provides 625ma or 1.2A, individual LED current = array current/6.



Schematic of the LED driver/constant current source used with the 36 LED array



Rob designed 3D parts for mounting the 36 LED array on his receiver set up to create a simple optical transceiver.

The 36 LEDs are the Osram model SFH4550 high power narrow beam width LEDs.

KOXL Receiver with 36 LED array mounted in the 1 inch square aluminum boom

Chart showing the comparison of the “on” power of the various transmitter configuration. Results taken at a distance of 2 meters.

K0XL Red LED Flashlight	600 microwatts	50% duty cycle	500 mA
WF0T IR 12 LED array	2600 microwatts	50% duty cycle	200 mA
K0XL Red LED Flashlight w/Fresnel lens	12400 microwatts	50% duty cycle	500mA

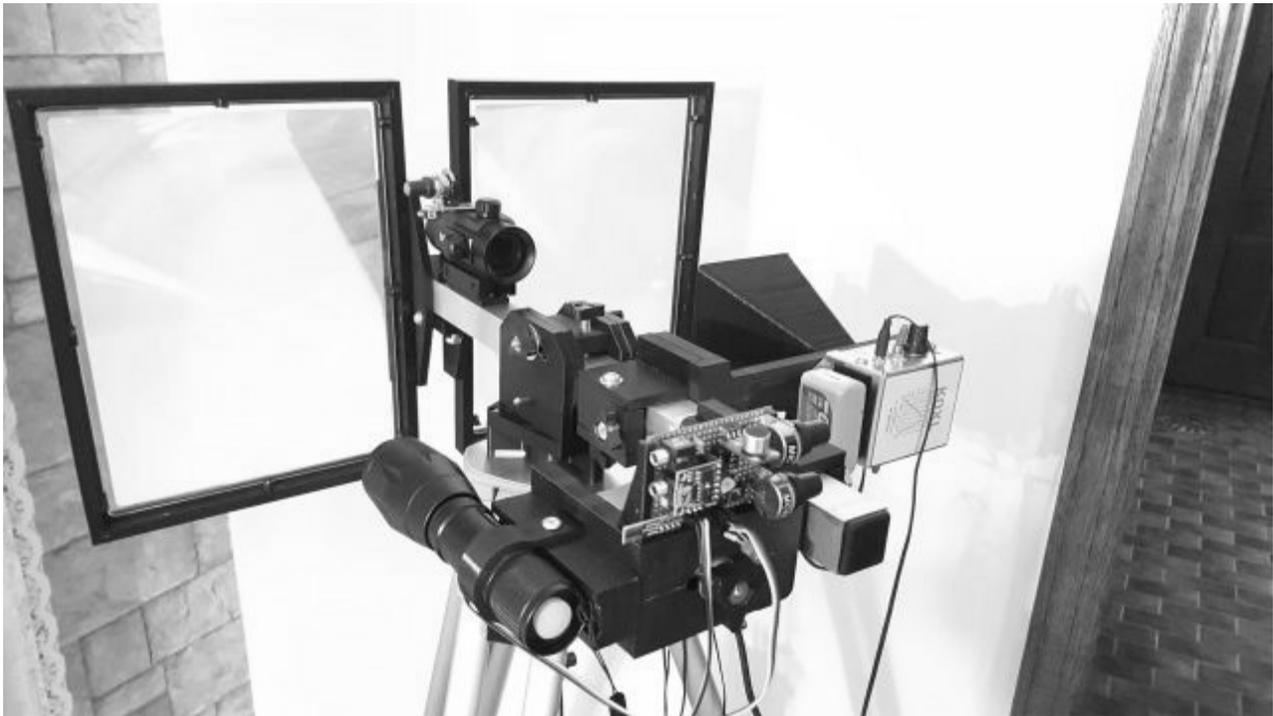


3-watt Flashlight transmitter lighting up the Sartell water tower from 1050 ft.

At this site we tested IR vs. Red light. Even though the Red light has about five times the light power over the 12 LED transmitter, the IR light won. The BPW34 sensor is only 67% efficient at red light and almost 100% efficient at 850nm (IR) which is probably the reason for the result. Rob was sending music to the tower and I was able to hear it just fine until I sent my IR light beam to the tower. With less light power the IR signal blocked out the red light signal so only the music I was playing was audible.

This test taught us that when bouncing light off of an object for communication, we need to aim at different points. When I moved my IR beam to the side of the big red spot, we were able to distinguish between the two signals. Peaking the receivers at the other’s transmitted light beam allowed us to hear each other’s signal with very little QRM from our own light beam.

This will be something we need to be aware of when we start cloud bounce communication testing.



Rob K0XL's Optical transceiver set-up

This is what happens when you can print your own parts! It's really an innovative idea, use the rail mount system to mount an optical transmitter on one side and the optical receiver on the other. Included in this picture are the transmitter mount, his ATtiny85 based modulator, a red-dot aiming device, a tilt mount for the tripod, his design for a lens holder and the lens clips he uses with his frame – along with the other components discussed earlier.

I have all of my 3D printed part files on my blog: wf0t.blogspot.com. I created a post in September called MUD 2019, where you will find the links.

Rob created a Google site where you can download all our 3D files as well as; more pictures, schematics, code used for our modulators, and more detail of some of the testing that's been done.

The link is: sites.google.com/view/k0xl

Note: 3D printer settings vary by manufacturer. I use .3mm layers with 20% infill with a .4mm print nozzle. I print with 3 layers for the top and bottom surface, with raft and support. Prints take from 11 minutes to about 2.5 hours. For the tripod mount and the receiver 3-axis mount I print these parts at 50% and 65% fill. Each print uses between 6 grams to 70 grams of PLA filament. The aluminum rail is 1-inch square.

73,

Warren WF0T & Rob K0XL

Small Dish Portable EME

By Rex Moncur, VK7MO

The keys to "Small dish Portable EME" are to:

- Use 10 GHz to get best system performance with small dishes.
- Use digital modes.
- Remove as many variables as possible (such as frequency, pointing and timing).
- Understand other variables such as libration spreading, lunar degradation, moon noise and absorption.
- Take advantage of "good" conditions, noting that there can be up to 10 dB and more variation between "good" and "poor" conditions.

Examples of what can be achieved at 10 GHz are:

- A four foot dish with 50 watts can work a similar station under most conditions.
- A two foot dish with 30 watts can work can work the larger home stations.

PROPOPAGATION LOSS

EME propagation loss increases by 20 dB for every 10 times increase in frequency such that:

144 MHz 253 dB

1296 MHz 271 dB

10,368 MHz 290 dB

37 dB more loss in going from 144 MHz to 10 GHz – so why go up in frequency?

ANTENNA GAIN

For the same size dish (or same capture area) the antenna gain goes up by 20 dB for every 10 times increase in frequency and thus fully compensates for the propagation losses. But as there are two antennas we gain 20 dB at each end for each 10 times increase in frequency. In the case of 144 to 10,368 MHz the overall gain is around 37 dB in going to 10 GHz. The result is we can use much smaller antennas that are suitable for portable operations. While 24 GHz provides another 7 dB antenna gain, it suffers from much higher atmospheric absorption, it is more difficult to generate useful power, has higher spreading and poorer receive noise figures - accordingly 10 GHz is considered the "sweet spot" for small dish portable EME.

TRANSMITTER POWER

In a portable situation it is difficult to generate more than about 50 watts at 10 GHz compared to much higher levels that can be generally used at lower frequencies - so in terms of overall system performance we don't get the full advantage of the higher antenna gain in going up to 10 GHz.

SYSTEM NOISE

In general system noise is lower at higher frequencies due to much lower external noise off-setting a small increase in the LNA noise figure. However, the higher gain and narrower beam-width of a 10 GHz antenna means it picks up more moon noise. For example, for a 10 foot dish receiving station this costs about 2 dB compared to only about 0.2 dB for a small 2 foot 6 inch dish. But even so 10 GHz is generally in front overall, in particular compared to the high noise levels in suburban environments at 144 MHz.

LUNAR DEGRADATION

As the moon moves around the Earth in an elliptical orbit its distance varies such that the propagation loss varies by up to 2.3 dB.

LIBRATION SPREADING

Lunar Libration spreads the signal in proportion to frequency and thus the effects of libration are 72 times worse at 10 GHz than 144 MHz. Typical spreading is around 100 Hz at 10 GHz and less than 1 Hz at 144 MHz. Spreading can, however, reach up to 200 Hz or more on 10 GHz and as shown in Fig 1 a change in spreading from 5 Hz to 200 Hz costs about 7 dB with the QRA64 mode.

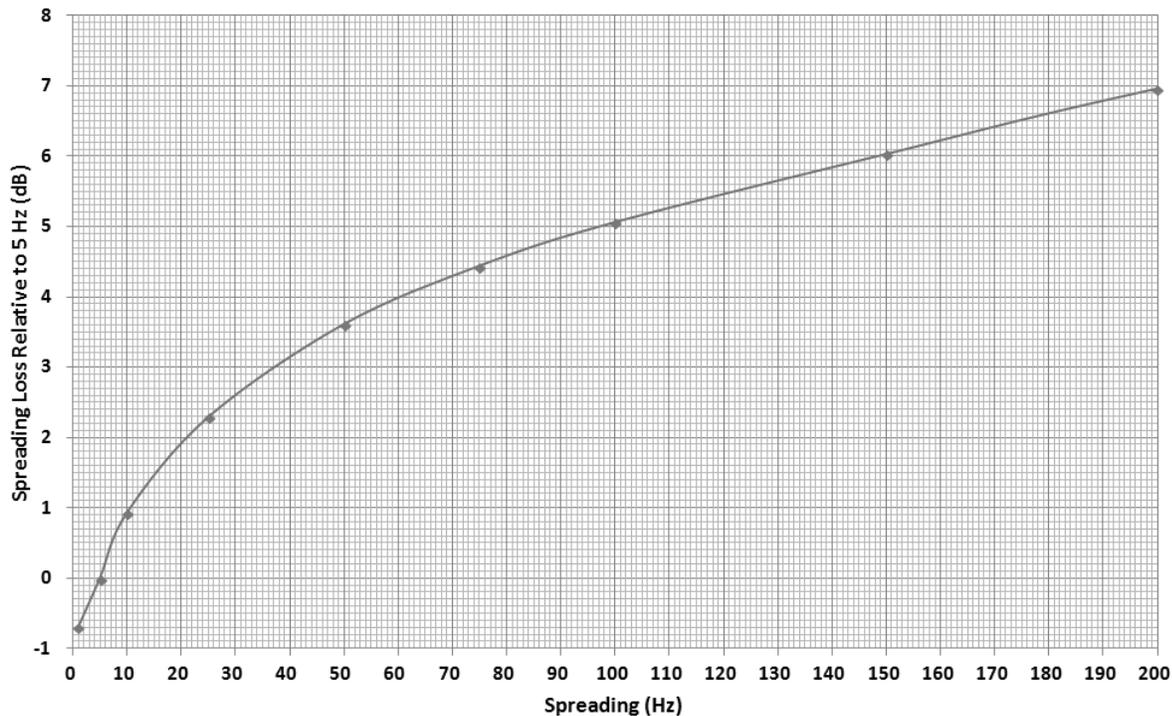


Fig 1: Libration Spreading Loss derives from DUBUS Paper by VK7MO and G3WDG⁽¹⁾

The good news is that there are times when the libration spreading drops to very low levels and full advantage can be taken of narrow bin-width sub-modes such that QRA64A. Such times only occur once per Lunar month but are the times one should choose for QRP EME (for example 5 watts with a

2' 6" dish). The Moonsked program by the late GM4JJJ shows both the spreading and lunar degradation so one can select the best time for QRP EME.

Libration spreading gets to its lowest values when stations are in close to antipodean positions. Thus a good option for QRP EME is from the East Coast of the USA to the South-West of Australia.

The following shows an example of the spreading over a month between VK7MO and G3WDG. In this case the libration spreading got down to 3 Hz on 11 April 2016 and thus there was essentially no libration loss.

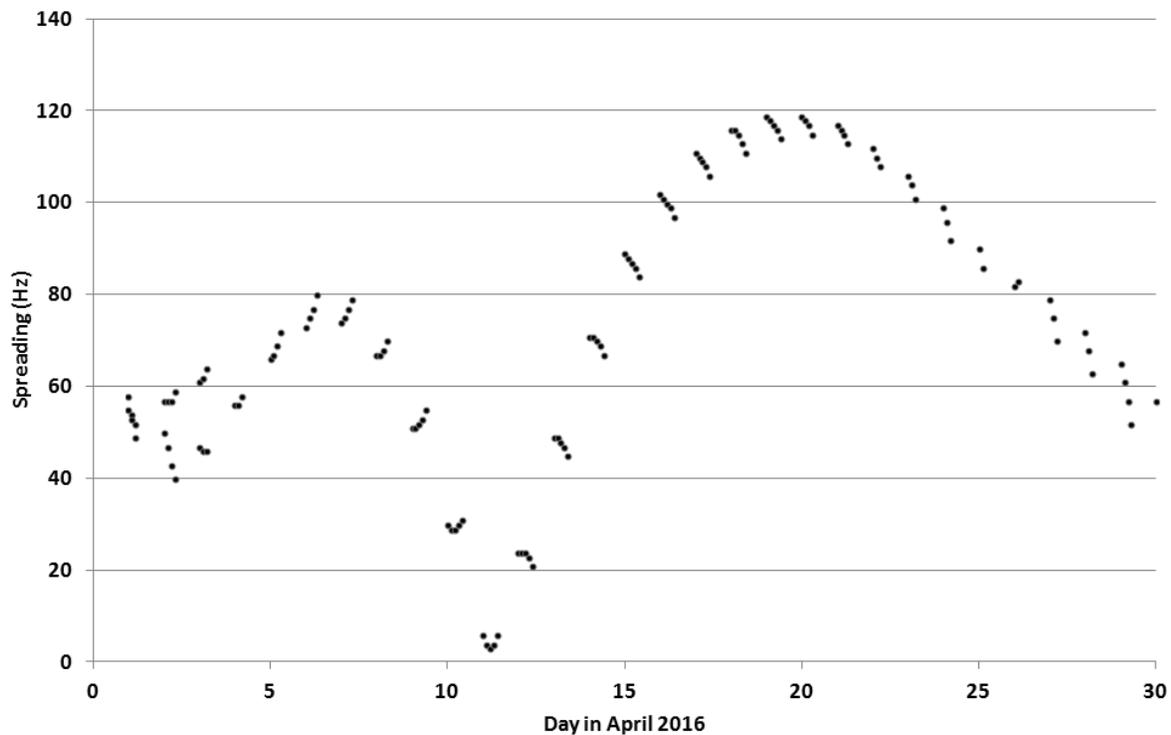


Fig 2: Example of Low Libration spreading.

ABSORPTION

Atmospheric absorption starts to become an issue at 10 GHz but is only significant when the Lunar elevation is below 5 degrees. Fig 3 shows that absorption can typically cost about 1 dB at 5 degrees and 5 dB at the horizon in temperate regions. In tropical regions it can increase by a factor of 3 or 4 times. At 24 GHz atmospheric absorption is about 10 times worse which rules it out for operations near the horizon and in particular in the tropics.

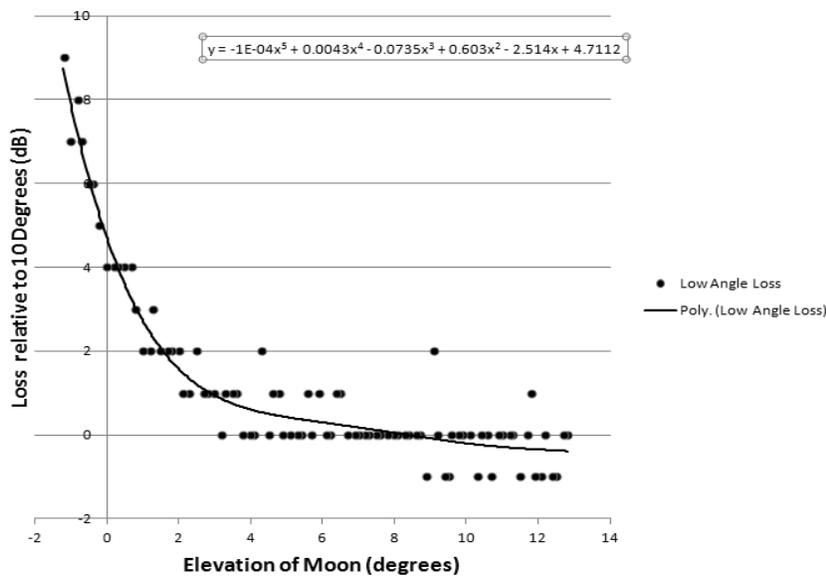


Fig 5: Loss due to absorption at 10 GHz measured by VK7MO from G3WDG⁽²⁾. (Precipitable water at the receiving end was 20 mm).

URLs to find forecasts of Precipitable Water are as follows:

North America: <http://wxmaps.org/pix/nam.pw>

Australia: <http://wxmaps.org/pix/aus.pw>

Europe: <http://wxmaps.org/pix/euro.pw>

As a rough guide the absorption loss (includes both attenuation and increased noise due to attenuation) at the horizon in dB is around one quarter of the Precipitable Water (PW) in mm. PW is typically around 40 to 60 mm in the tropics and 10 to 30 mm in temperate regions. This means that small dish operations over long distances where the elevation angles are close to the horizon are much more difficult in tropical areas.

GROUND NOISE

Ground noise can be an issue at 10 GHz when beaming close to the horizon as is necessary to achieve very long distances. As shown in Fig 5 below when beaming over the sea the ground noise continues to increase on Vertical polarization but does not on horizontal polarization. It is postulated that this is because Horizontal Polarization reflects rather than emits radiation from the water, where-as Vertical Polarization does not reflect and thus receives the emitted noise from the water. This feature can be taken advantage of to improve system performance of long distance EME such as breaking World Records ⁽²⁾.

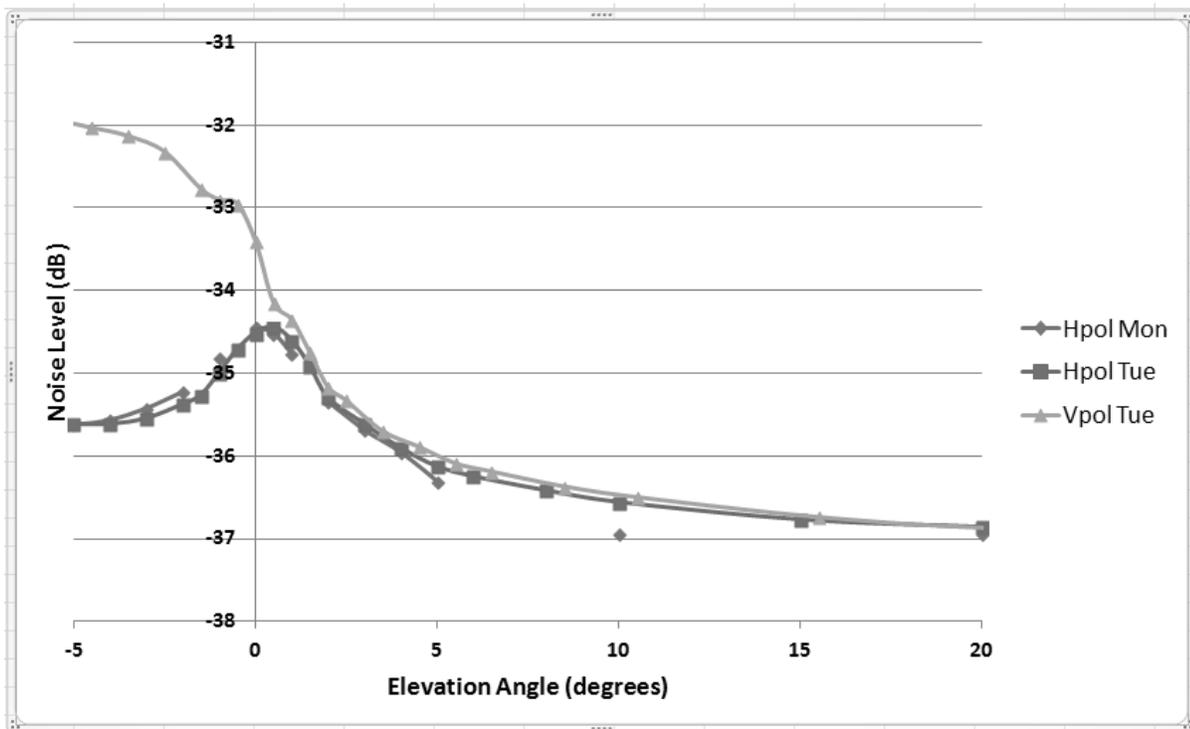


Fig 5: Ground and Absorption Noise measured over the sea by VK7MO with the assistance of VK7WLH.

MODE

The QRA64D sub-mode in WSJT-X, developed by Nico Palermo, IV3NWV⁽³⁾, and Joe Taylor, K1JT⁽⁴⁾ is generally used at 10 GHz and gains 10 to 16 dB on CW. It is more sensitive than JT4f up to 130 Hz spreading⁽¹⁾.

VARIATION OF CONDITIONS

As seen above libration spreading can produce variations of up to 7 dB, Luna degradation up to 2.3 dB, and Moon noise up to 2 dB. So overall one can have variations between "good" and "poor" conditions of up to around 10 dB. Below 5 degrees elevation absorption can add significant additional loss as indicated above.

REMOVING VARIABLES

Frequency and Doppler: At 10 GHz Doppler shift varies by over 40 KHz and can change at up to 1 Hz/second. Accordingly, it is necessary to use automatic Doppler correction such as provided by WSJT-X. In addition the absolute frequency needs to be not only stable to within a few Hz over a transmit cycle but to gain maximum sensitivity it is desirable to have the absolute frequency accurate to within 10 Hz so that one can use a narrow tolerance (10 or 20 Hz) on WSJT-X to reduce false syncs to a minimum. To meet these requirements it is necessary to GPSDO lock the transverter and preferably the IF radio.

Tracking the Moon: Small dishes do not pick up enough Moon noise for tracking and one must have a system of tracking to better than 0.5 degrees in both Azimuth and Elevation. An inclinometer is adequate for elevation but Azimuth requires some known reference. The key is to use a rifle-scope which has been previously calibrated to the dish on Sun noise. If the Moon is visible one can track the dish optically or at least set the Azimuth calibration when the Moon appears through clouds. The Sun can also be used for calibration if it is above the horizon and visible (but you must provide protection from eye damage such as by covering the rifle-scope with a No 13 welding slide). If it is cloudy one can use some feature such as a tower or house that is visible and identifiable on Google Earth. But there are some situations where nothing is available for an Azimuth reference in which case VK3HZ has developed a differential GPS system that can give accurate Azimuth bearings to better than 0.1 degrees on a 5 metre baseline⁽⁵⁾.

Polarization: At 10 GHz, circular polarization is rare and most stations use linear polarization. Linear to linear polarization has been shown experimentally by G3WDG ⁽⁶⁾ to have an advantage of about 1.5 dB over circular to circular due to the characteristics of lunar reflection, it is easier to set up and small portable stations are also used for terrestrial where linear polarization is the norm. Horizontal linear polarization also has advantages in lower ground noise over the sea⁽²⁾. The general rule for DXpeditions is that the portable station decides on the polarization and other stations who wish to work them adjust their polarization to suit. While I generally use Horizontal Polarization I can adjust polarization by the use of brackets to mount the dish for Vertical Polarization and +/- 45 degrees.

Timing: WSJT-X requires timing to be accurate to better than 1 second to reliably decode. Internet time will often not be available in the field. GPS timing is the solution for portable operations and can be implemented in-expensively with a USB GPS receiver and a program called NMEATime to give results to better than 0.1 second ⁽⁷⁾.

POWER AMPLIFIERS

The cost of Power Amplifiers is the biggest negative in going to 10 GHz EME. TWTs are available on the surplus market but require very high voltages, which raises questions of arcing over due to humidity and rain. Safety when operating in the field at night is another concern. 60 watt Solid State Amplifiers are available from Khune, DB6NT at about US\$ 3700 and 30 watt PAs for about half that price. The majority of home stations run at least 50 watts and given that they generally have large dishes and must cope with around 2 dB more moon noise it is preferable that the portable station run at least 50 watts. Never-the-less if one is prepared to wait for "good" conditions one can make many contacts with the larger stations with 10 to 20 watts and a 2 foot 6 inch dish.

PRIME FOCUS OR OFFSET DISH

An offset dish should have improved performance due to less waveguide loss, less blocking and lower noise as it's feed points to the sky. The problem with offset dishes is that they are not normally designed to beam at or near the horizon and the feed, pre-amp, waveguide switch and heavy PA are all well out in front making it difficult to maintain mechanical balance on a tripod. While you can mount

a small offset dish such that the bottom is fully above the tripod this becomes problematic for anything larger than a 2 foot dish. The big advantage of a prime focus dish is that you can use a Shepherd's crook waveguide feed and mount all the heavy gear behind the dish to give mechanical balance on a tripod - thus a prime focus dish is much preferred for small dish portable EME.

AUTOMATIC TRACKING

I have never used automatic tracking for portable operations. Primarily because this is one more thing to carry, set up and go wrong, but also because manual tracking is easy to operate during the transmit and receive periods on WSJT.

PHYSICAL ISSUES

A 2 foot 6 inch dish fully assembled and ready to go can be carried in the back of an SUV - this has the advantage that it can be set up very quickly and I have for example worked 3 or 4 stations at each of 4 separate grid locators on a single Moon Pass. A four foot dish gives a very useful 4 dB more gain but needs to be carried on the roof of a vehicle and assembled on site. Above four foot becomes very difficult to assemble by one person and is more suitable for fixed location dxpeditions rather than collecting multiple grid locators.

DL0SHF BEACON

The DL0SHF Beacon runs the QRA64D sub-mode on 10368.025 MHz with 40 watts to a 24 foot dish and Vertical Polarization. It is an excellent way of testing small dish systems or small horns. Tune to 10368.024 MHz and run "full Doppler" on WSJT-X.

WORLD DISTANCE RECORDS

Portable to Portable operations can be used to find locations with close to zero elevation take-offs to achieve World Record distances as below:

VK7MO (4 foot dish and 60 watts) to the late WA3LBI (8 foot dish and 150 watts), 18725 km ⁽⁸⁾

ZL/VK7MO (4 foot dish and 60 watts) to G3WDG (4 foot dish and 75 watts), 19105 km ⁽²⁾

4 foot dishes and 60 watts combined with "good" conditions are required to overcome absorption losses at the low elevations required for World record distance contacts.

SYSTEM RELIABILITY

On a typical portable DXpedition I might travel over 10,000 miles - much of it on rough dusty roads. At each new grid locator it is necessary to find a suitable location - preferably with a take-off to both moon-rise for North America and moon- set to Europe. The equipment must be unpacked, set-up and many cables need to be connected (there is a right place for everything and it must all be re-packed exactly as it was packed). An accurate azimuth reference must be determined. The equipment must then be set up and tested well in advance so there is time to resolve any issues. I carry as much

redundancy as possible - two IF radios, two GPSDO's and two computers and for long trips a back-up 10 GHz system with a smaller dish. An independent GPSDO locked weak signal source is very useful for system testing as is the simple test of cold sky to ground - PA current is a good means to check that power is coming out. Operations will often have to be conducted at night without upsetting other travelers who are sleeping. If you are to maintain the interest of hams around the World you must come on each time you say you will. This means not only that the equipment is reliable but being in position when you say you will be - so things like sleep and meals are secondary. Internet connectivity will often not be available.

OPTIONS

Fig 4: compares the power required for "good" and "poor" conditions with various antenna sizes following:

Dish Size	Best Case (watts)	Worst Case (watts)
2' 6" to 4 foot	15	Too much
2' 6" to 10 foot	3	30
4 foot to 4 foot	5	50
4 foot to 10 foot	1	10

PRACTICAL RESULTS

- 4 x 5 inch aperture horn (equivalent to a 5 inch dish) worked G3WDG (10 foot dish and 75 watts) under "very good" conditions.
- 2 foot back-up dish and 60 watts, worked several grid locators to stations with 8 foot dishes and above.
- 2 foot six inch dish and 50 watts, activated over 100 grid locators around Australia.
- 4 foot cut-up petal dish and 60 watts, worked G3WDG 4 foot dish and 75 watts for the current World 10 GHz EME record of 19105 km.

ACKNOWLEDGEMENTS

To Charlie Suckling, G3WDG, who has been part of the many experiments that have led to my being able to understand and improve small dish EME operations. To Glen English, VK1XX, who developed the initial Doppler correction system that is at the heart of this work and his contributions to understanding absorption and ground noise. Joe Taylor, K1JT, and his development team for the development of WSJT-X including QRA64 which was formulated by Nico Palermo, IV3NWV. To OK1KIR, W5LUA, G3WDG, G4KGC, OK2AQ, OZ1LPR, the Late WA3LBI and numerous others who have been at the other end of my portable contacts. To ZL3RC and ZL3FJ in supporting the current World Record. To VK7WLH, VK7ZBX and VK7ZFC who supported tests in Tasmania. To VK7WLH for reviewing this paper.

CONCLUSIONS

A 4 foot dish is physically viable for portable operations and with 50 watts can work a similar station under even the worst conditions. A 2 foot 6 inch dish can be carried set-up inside an SUV or operated from a small balcony and will allow you to work most home stations under average conditions.

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VHF/UHF/Microwave High-Power LDMOS Amplifiers

Greg McIntire, AA5C

Vacuum tubes were the technology mainstay for legal limit amateur high-power HF, VHF, UHF, and microwave amplifiers for many years. During the mid-1980s I built kilowatt amplifiers for 6m, 2m, and 70cm using pairs of 4X150A (6m) and 4CX250B tubes (2m and 70cm). A modified surplus Federal Aviation Administration (FAA) cavity with a single 4CX250R was my amplifier for 1.25 m operation. These amps served me well for decades in CW and SSB service. The use of digital modes has become ubiquitous, however, and the high duty cycle of these modes required backing off on the output power. The need to limit on-time duration became particularly obvious when the Teflon insulating tube in which the high voltage supply line is run for the FAA cavity amp melted resulting in a high voltage short during a meteor scatter run on 222 MHz. This “exciting” event reinforced the limitations of my amps. Since I want to be able run high power for relatively long on periods, e.g., 1 minute for JT65, I decided it was time for a technology refresh. My goal was to replace the four tube amplifiers in my VHF/UHF rack with solid state amps capable of running JT65 with at least 1KW output on each band and to replace my traveling wave tube amplifier (TWTA) on 2.3 GHz.

I started doing some research and the Laterally Diffused Metal Oxide Semiconductor (LDMOS) power FET technology seemed to be the best to turn to for high power VHF/UHF amplifiers. The cell phone industry helped drive the development of this technology in the late 1990s and the technology has continued to advance to this day. Early devices, during the 1990s, operated with a 28 VDC drain supply. That increased to 50 VDC during the 2000s and early 2010s. Now devices operating with a 65 VDC drain supply are available. One benefit of the higher drain voltage is the more manageable drain current compared to lower voltage devices running comparable output power. Another is that the matching networks are not as difficult to design because the output impedances aren't as low. LDMOS has excellent linear performance when operated Class AB and this makes it possible to built linear amplifiers with high gain and good efficiency at a reasonable cost. The high peak power capability of the devices makes them suitable for the multicarrier CDMA systems often used in the cell phone industry. Another advantage of modern LDMOS devices is that they are very tolerant of high output VSWR with a specification of surviving up to 65:1 VSWR being common (for a 100 us pulse, 20% duty cycle). The high gain helps reduce the number of stages required to achieve a desired output power in a system. Another factor that simplifies the overall system is that they are enhancement mode devices so a single positive supply can be used with a step down regulator or other arrangement for the gate bias supply. Most devices are dual to facilitate push-pull operation. The Ampleon BLF188XR is in a typical dual LDMOS package (Figure 1).

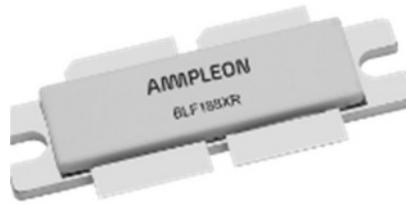


Figure 1 Ampleon BLF188XR Dual LDMOS Transistor

The primary disadvantage of LDMOS devices is that they are very sensitive to excessive gate voltage or drive power. I took out several devices in one 2.3 GHz amplifier because the potentiometer I was using to set the gate bias (from a regulated 5 VDC supply) turned out to be noisy. Some radios put out an RF glitch when first going into the transmit mode and this can also take them out. The typical dynamic means of controlling drive, automatic gain control (AGC), is woefully too slow to protect devices. Check your driver before applying input power.

A few of the key specifications for some common VHF/UHF/microwave LDMOS transistors are shown in Table 1. The high power capability, high gain, and high operating frequency of some of the more current devices are noteworthy. Also note that current devices are generally available from distributors like Digi-Key and Mouser. I've had good luck finding some of the obsolete parts on eBay.

Device	Manufacturer	Drain Supply	Gain (dB)	Pout @ Frequency	Sources
MRF151G	MaCOM	50V	14	150W/FET up to 175 MHz (300W total)	Mouser, eBay
MRF1K50	NXP	50V	23	1.5 KW total@ 144 MHz	Mouser
MRFX1K80	NXP	65V	24	1.8 KW total@ 144 MHz	Mouser
BLF578XR	Ampleon (NXP)	50V	23.5	1.4 KW @ 225 MHz	Digi-Key
BLF184XP	Ampleon (NXP)	50V	23.9	700 W @ 432 MHz	Digi-Key
MRF13750	NXP	50V	17.2 @ 1.3 GHz	750W @ 1.3 GHz	Mouser
MRF19125	NXP (Freescale)	26V	13.6	125W @ 2 GHz	Obsolete (eBay)

Table 1 Key Parameters of a Few LDMOS Power Transistors

The popularity of the technology is evidenced by the number of LDMOS transistor manufacturers. Consolidations, buyouts, and mergers make the list dynamic. It includes: NXP Semiconductors, Ampleon, Infineon Technologies, Qorvo (RFMD), STMicroelectronics, Freescale Semiconductors, TSMC, LFoundry, Tower Semiconductor, Globalfoundries, Vanguard International Semiconductor Corporation, SMIC, MK Semiconductors, and Polyfet.

There are varied approaches one can take in getting high power LDMOS amplifier capability in the shack. Several hams, including Rob McCance, N4GA (Reference 1) and Jim Klitzing, W6PQL have designed and built LDMOS-based amplifiers. Jim Klitzing, W6PQL wrote an article on one 2m KW LDMOS amp he built. The article was published in the October 2012 issue of QST (Reference 2). He built it using an MRFE6VP62K25H LMDOS transistor and notes the cost advantage compared to tubes to generate comparable power in his article. Jim offers turn-key amplifiers and when I was putting together my 2m EME station in 2018 I chose a custom 1.5 KW amplifier from him. Jim also offers the basic parts and RF input/output splitter/combiner boards, copper heat spreaders, and so forth so that you can build your own RF deck from scratch at whatever level you want to start from. For 432 MHz I got a 1 KW deck from Jim and built up the amplifier with a few other pieces he offers, e.g., a control board, filter/dual direction coupler, and high current FET switch for the drain. There are still a number of surplus commercial LDMOS amplifier decks available from different sources. The switchover from analog to digital television resulted in a lot of LDMOS 1 KW and 1.5 KW amplifier decks becoming available. Most notable of these are amplifiers built by Harris and Larcan. I'm most familiar with the Larcan amplifiers of which there are 1 KW amplifier decks with 4 dual LDMOS transistors and 1.5 KW decks with 6 dual LDMOS transistors. One set of circuit boards covers the 54 to 88 MHz range: 54-72 MHz and 76-88 MHz depending on capacitor selection and strapping of printed inductors. These are referred to as the VHF LO and VHF Lo/Hi amplifiers, respectively. Another set of circuit boards is used for the VHF Hi Band amplifier that is designed to cover 174-216 MHz. In my case, I started with two 1.5 KW Larcan decks. One deck is a VHF Lo/Hi deck (Figure 2) Figure 2 Larcan VHF Lo/Hi 1.5 KW Deck and the other a 1.5 KW VHF Hi deck (Figure 3). Many excellent articles on modifying these amplifiers have been written. The Minute Man Repeater Association web site (Reference 3) is a particularly good repository of schematics, modification articles, and other information for Larcan amplifiers.

Careful study of the tables in the Larcan schematics and review of an article by David Olean, K1WHS (Reference 4) led me to an approach of changing 36 ATC capacitors and removing the straps on 12 strip line inductors to get the VHF Lo/Hi deck to efficiently operate on 50 MHz. The VHF Lo version will go down to 54 MHz but further modifications are needed to get efficient operation at 50 MHz. The capacitor modifications were a combination of changing out ATC caps and stacking (paralleling) ATC caps to achieve the desired value. Brian Justin, WA1ZMS has taken the VHF Lo/Hi deck up in frequency and shows how to put it on 2 meters in one of his articles (Reference 5). The Larcan Hi band amplifier works directly on 222 MHz so matching circuit modifications were not required. The other modification I made to the basic RF deck is to replace the input connector with a conventional female BNC connector and the output connector with a conventional N connector. This was relatively straightforward, requiring the fabrication of a few adapter plates. The stock Larcan amplifiers have input and output RF connectors that are designed for hot switching the decks by sliding them into a powered rack but they are not easily adapted to for use with common connector types. DC power is supplied to the deck through pins on the back side. I found that different sized terminal lugs fit the different power pins. The crimp side of the lug made a good compression fit on the power pin leaving the hole or spade end of the lug available as an attachment point.

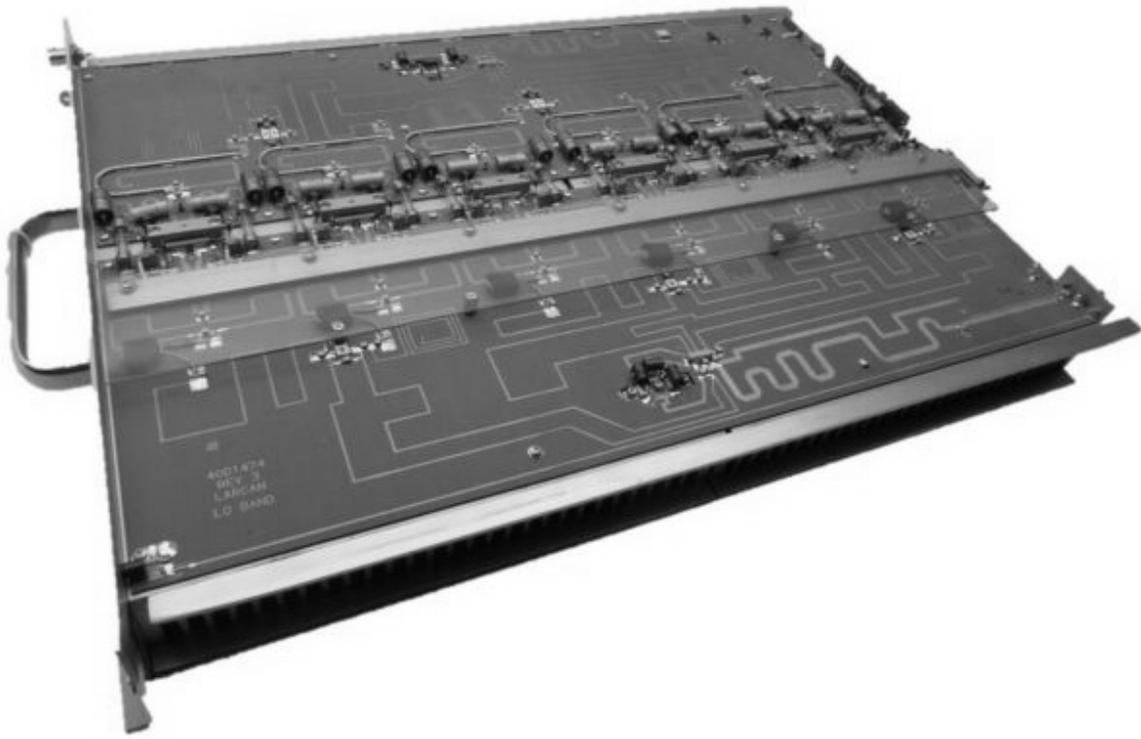


Figure 2 Larcam VHF Lo/Hi 1.5 KW Deck

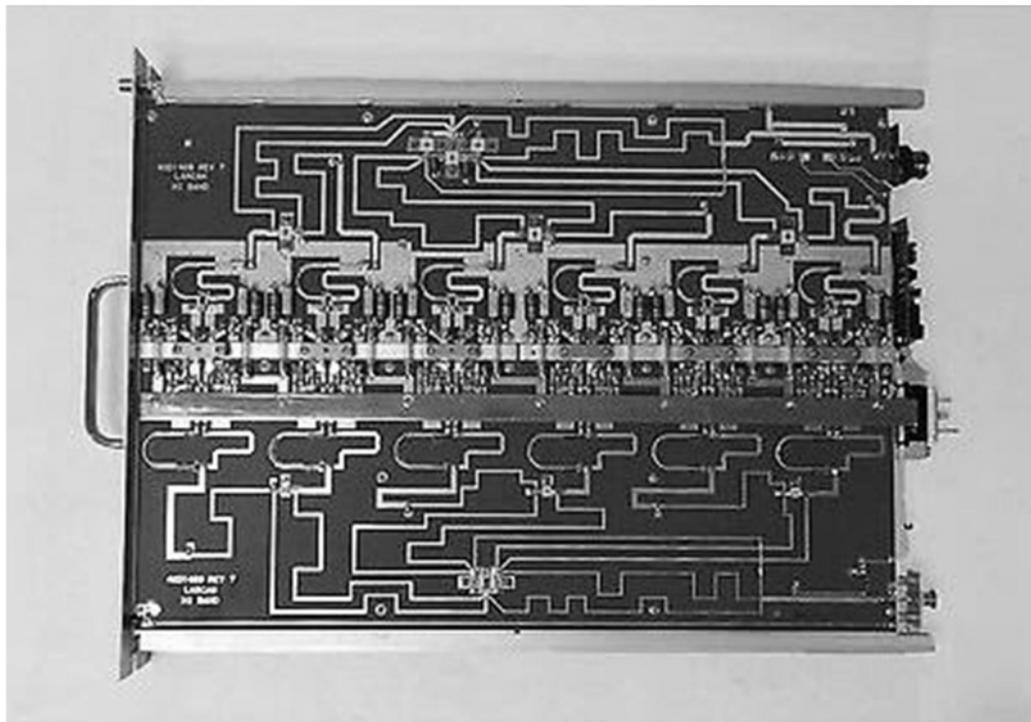


Figure 3 Larcam VHF Hi 1.5 KW RF Deck

Surplus “blade” computer server power supplies are a great main component of a power supply for the LDMOS amplifiers. The supplies are typically rated at 51.4V at 57A. They operate from a 220 VAC supply and one or two jumpers can be added to make them operate outside the racks they normally plug into. Rob McCance, N4GA (Reference 1) shows one way to jumper the units so that they turn on outside their intended rack connections. Other circuits are needed to safely operate the amplifier deck with the power supply. Another good supply I’ve recently become aware of is the Eltek 48V/3000 HE Flatpak. The Larcans decks have 2 sets of power pins. One set (the longer ones) are for supplying 50 VDC to the step-start circuit. This basically allows the 12 large electrolytic capacitors to charge through a limiting power resistor before direct 50 VDC power is applied. This helps avoid blowing the fuses that are in series with the drain of each device. The second set of pins, the shorter and larger diameter ones, supply 50 VDC at high current to each stage via a bus bar. The arrangement of the power supply pins is a clever means of mechanically accomplishing the sequencing as the amplifier is plugged into a rack but without a rack, an external sequencer is needed. I built up two sequencer boards, one for the 50-54 MHz amp and another for the 222 MHz amplifier. These sequencers are based on the venerable W5LUA design that uses an LM324. The sequence I use for going from receive to transmit is:

1st: switch the T/R relays to the transmit state

2nd: apply 50 VDC to the step-start circuit

3rd: apply 50 VDC to the full amplifier

The reverse sequence should be followed when going from transmit to receive.

One of the power supply trays I built is shown in Figure 4. The 50 VDC power supply is the long unit at the top of the picture. The tray includes 24 VDC and 12 VDC power supplies for T/R relays, control relays, and the sequencers. A 50A ammeter and shunt are included for monitoring current. LEDs on the front panel indicate the status of each power supply.

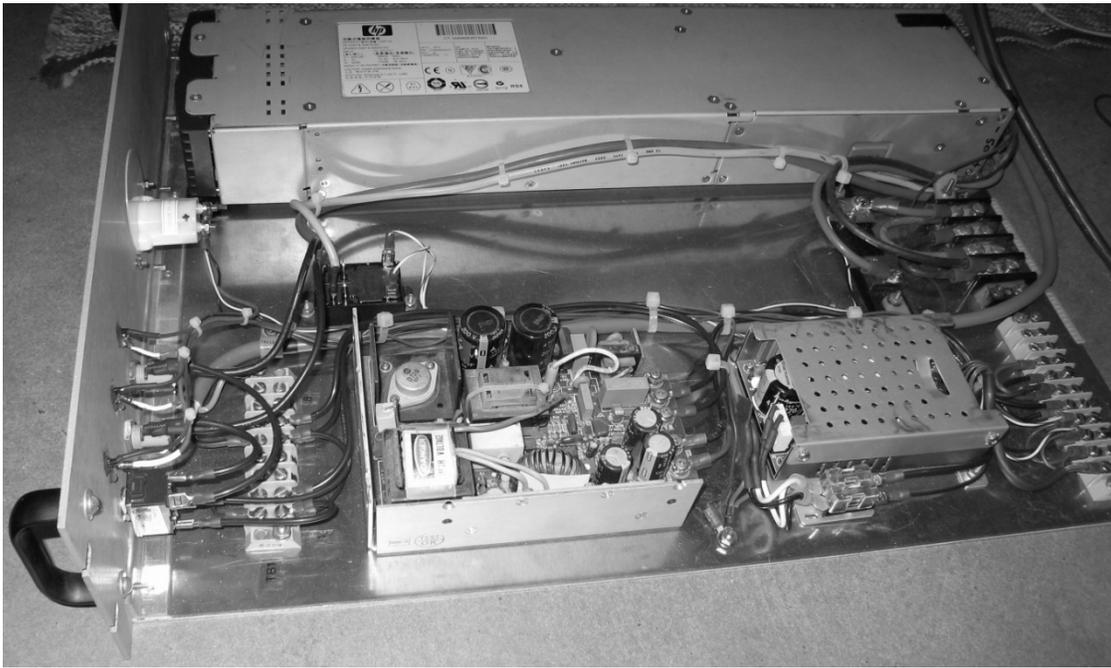


Figure 4 Power Supply Tray for High Power LDMOS Amplifiers

Each complete amplifier should include a circuit for shutting down the amp if the output VSWR is too high, typically over 1.5:1. The Larcan amplifiers include a control board located along the back side of the deck between the input and output connectors. This control board includes the current limiting resistor for charging the electrolytic capacitors and a VSWR monitoring circuit. The strip line couplers for detecting forward and reverse power are part of the two main RF circuit boards. The control board also feeds a front panel LED that illuminates when the VSWR is lower than 1.5:1 and output power is more than several hundred watts. I find this useful for a quick check when operating. The Ultimate Control Board designed by Jim Klitzing, W6PQL includes a shut down circuit for high VSWR. He also designed boards for the different bands that combine a low pass filter and forward and reverse power detectors. A front panel 10-LED bar graph indicator for forward power and another for reverse power fed by the detectors make good quick reference indicators during operation. The reverse power signal from the detector also feeds the control board high VSWR shut down circuit.

I used 0.125 inch thick aluminum sheet metal for the base plates and 6U (10.5 inch) aluminum rack panels to build up the amplifiers mechanically. Four 120 mm 54 CFM cooling fans are mounted to the underside (heat sink side) of each deck using $\frac{3}{4}$ inch aluminum angle. The sequencer and control relays and interconnect wiring are mounted on the base plate along with the input and output RF coaxial relays for T/R switching. Fuses, switches, and indicators are mounted on the front panel. An RF attenuator on the RF input to the amplifier is included. The attenuator is sized to make sure that full output power from the transceiver cannot overdrive the amplifier. A starting point for calculating each attenuator value is the maximum output power of the transceiver and the maximum gain of the LDMOS amplifier. Expected gain of each Larcan deck is listed in the manual for that version. I

measured the gain of the modified VHF Lo/Hi amp to be 20 dB at 50 MHz and that of the VHF Hi amplifier to be 16 dB at 222 MHz. Be cautious and start with higher attenuator values than you think you will ultimately need.

I pulled all of the tube amplifiers and power supplies from my VHF/UHF amplifier rack and installed the necessary supports and rack hardware to complete the installation. The final result is shown in Figure 5. The top half of the rack includes the 1.5 KW 6m and 1.25m amplifiers fed from a common power supply tray. The bottom half of the rack includes the 4U (7 inch) sized 1.5 KW 2m amplifier (W6PQL) and the 1 KW 70 cm (AA5C using a W6PQL RF deck) amplifier. These two amplifiers are fed by a second common power supply tray. A single power supply tray could have been used with good bussing of the 50V lines but this arrangement lets me switch supplies if I have a failure.

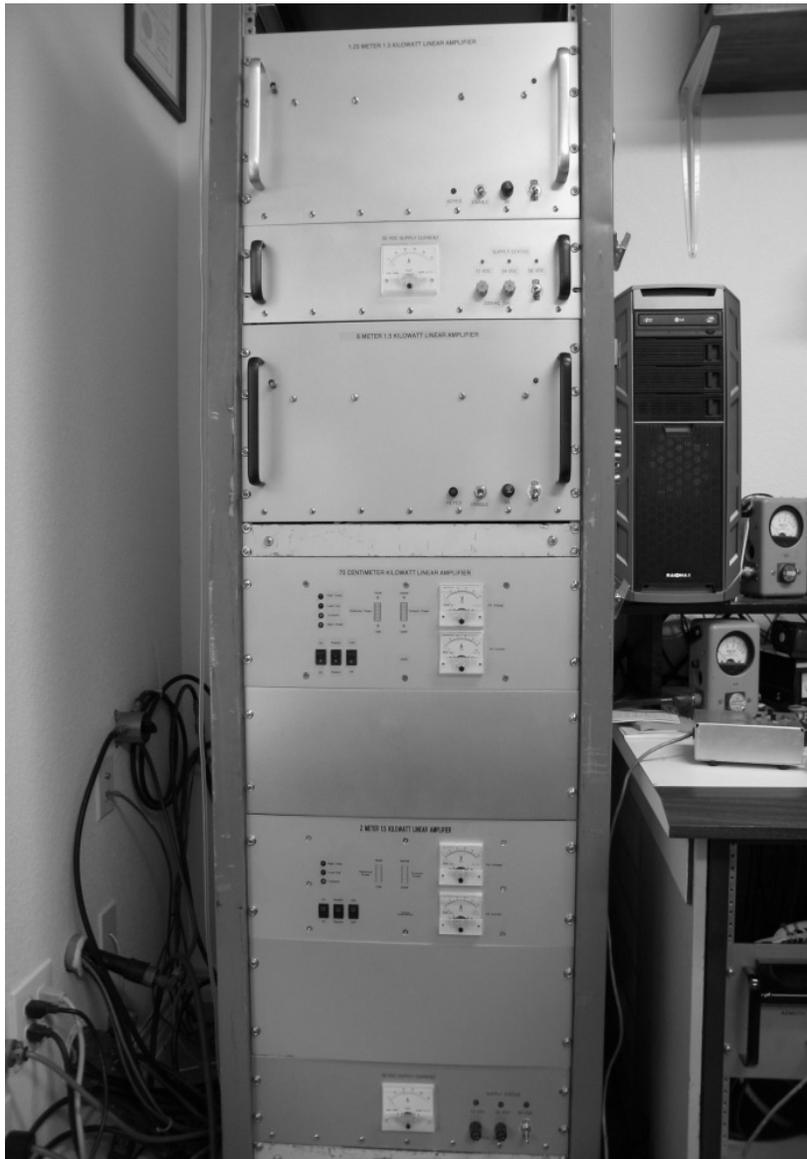


Figure 5 AA5C LDMOS KW Amplifier Rack

The output power versus operating frequency curve continues to increase as the LDMOS technology advances. Jim Klitzing, W6PQL has designed a 600 W 1296 MHz LDMOS amplifiers. The RF deck (Figure 6) is built around the MRF13750 LDMOS dual transistor. This amplifier produces 600 Watts output with 10 Watts drive.

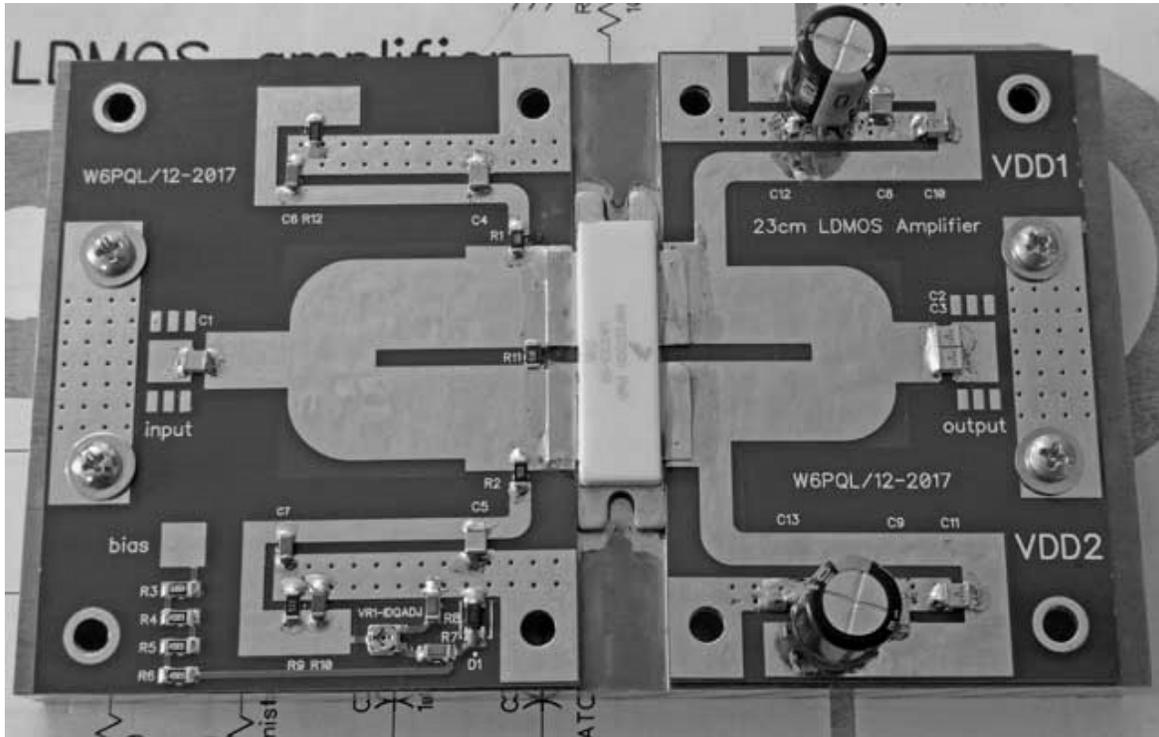


Figure 6 LDMOS 600 Watt 1296 MHz Amplifier Deck (photo courteous W6PQL)

My experience with LDMOS on the microwave bands has primarily been from adapting surplus commercial Personal Communications Service (PCS) amplifiers to the amateur bands. One PCS band, 824-894 MHz, is close to the 902 MHz amateur band and the other PCS band, 1850-1990 MHz is close to the 2.3 GHz amateur band. The surplus commercial amps tend to be a generation or two behind and generally run on 28 VDC with output power in the 100 Watt range. One amp I converted for 902 MHz yielded 125 Watts output with a few Watts drive. For 2304 MHz I converted an AML amplifier based on an article R.L Frey published in the 2006 Eastern VHF/UHF Conference Proceedings (Reference 6). Plans are to replace my old 25W TWTA with it once I design and build a suitable driver amp.

My objective has been to provide information and references for you to be able to start working with LDMOS transistors and amplifiers. LDMOS technology is rapidly replacing tube technology and solid state operation has some distinct advantages over tubes. The front panels of my solid state amplifiers are devoid of tuning controls and the instant operation is very convenient.

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Use of WR-28 Waveguide on 47 GHz

By Barry Malowanchuk VE4MA

Introduction

This paper originated as an unpublished presentation made at Microwave Update in 2004 when preparations were being made for the first 47 GHz EME attempts by VE4MA, W5LUA, AD6FP and RW3BP. Today with so many more amateurs preparing for 47 GHz operation, it is felt that although dated, the information is still of significance.

Waveguide is the standard transmission line used for the amateur 47 GHz band. Several waveguide sizes (WR-22 & WR-19) are designed for possible use at this frequency, and WR-15 will also work, however none of these are commonly available on the surplus markets and when found is rather expensive. Fortunately in 2019 Kuhne Electronic is now selling straight pieces of WR-19 waveguide and corresponding round waveguide flanges, so at least one easily accessible source exists. Several amateurs have been able to bend these pieces after heat treating them, so that some construction flexibility is gained.

There is however a large surplus supply of low cost WR-28 waveguide, designed for 26 to 40 GHz, and is known to work down at 24 GHz. WR-28 comes with round and square waveguide flanges, which are compatible with WR-22 round & square flanges and the round WR-19 flanges commonly seen on Kuhne Electronic modules. There has been however very little information about WR-28 behavior above 40 GHz and in particular at 47 GHz.

From previous work it is known that non-standard oversized waveguides operate with very low loss at 24 GHz, as long as straight runs are used. Bends in the H plane are believed to be susceptible to operating in higher modes as the operating frequency is increased beyond the designed range. This paper will cover the Insertion and Return Loss testing of a variety of WR-28 pieces as shown in Figure 1 to a WR-22 system. It will not however examine the losses associated with mating WR-28 waveguides to a WR-19 system but the findings with WR-22 testing suggests this can be quite low on a narrowband basis.

Waveguide Moding Problems Anticipated With WR28

The dominant TE_{10} propagation mode is dependent on the width of waveguide used at any particular frequency. The standard operating ranges are defined for the reliable operation of the TE_{10} propagation mode. If we operate out of the standard range the TE_{10} mode may continue to operate reliably as long as there are no discontinuities in the waveguide. So straight waveguide should be more reliable than if we have bends, twists, etc.

Referring to Figure 2, you will see the various fields associated with the dominant TE_{10} waveguide mode. The Cut-off Frequency (lower) is 21.1 GHz for WR-28.

In Figure 3 you will see the next higher waveguide mode TE_{11} . The cut-off Frequency for this mode in WR-28 is 47.2 GHz.....which is an obvious concern for 47 GHz operation!

Figure 4 shows another possible higher mode TE_{21} . The cut-off Frequency for this mode in WR-28 is 59.6 GHz so it's not a concern.



Figure 1 Assortment of WR-28 Waveguide Parts for Testing

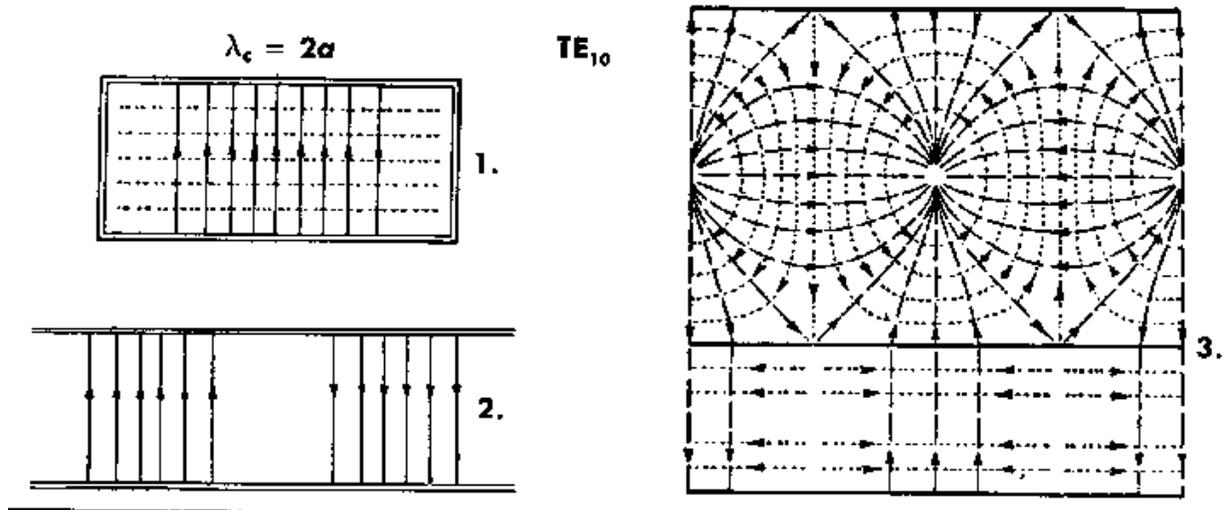


Figure 2 TE₁₀ Propagation Mode (Cut-off Frequency = 21.1 GHz)

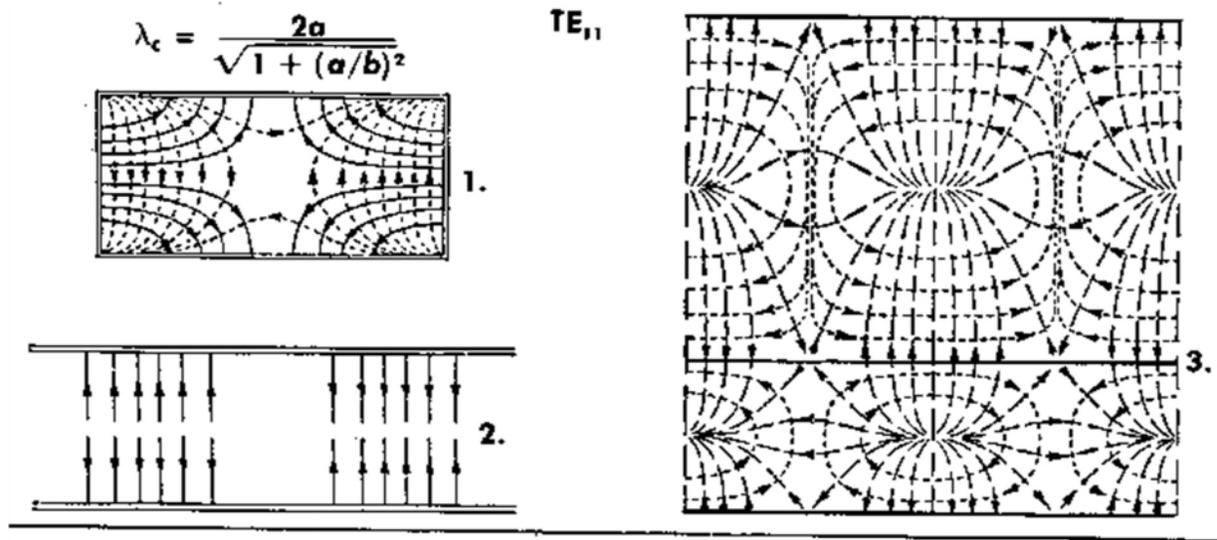


Figure 3 TE₁₁ Propagation Mode (Cut-off Frequency 47.2 GHz)

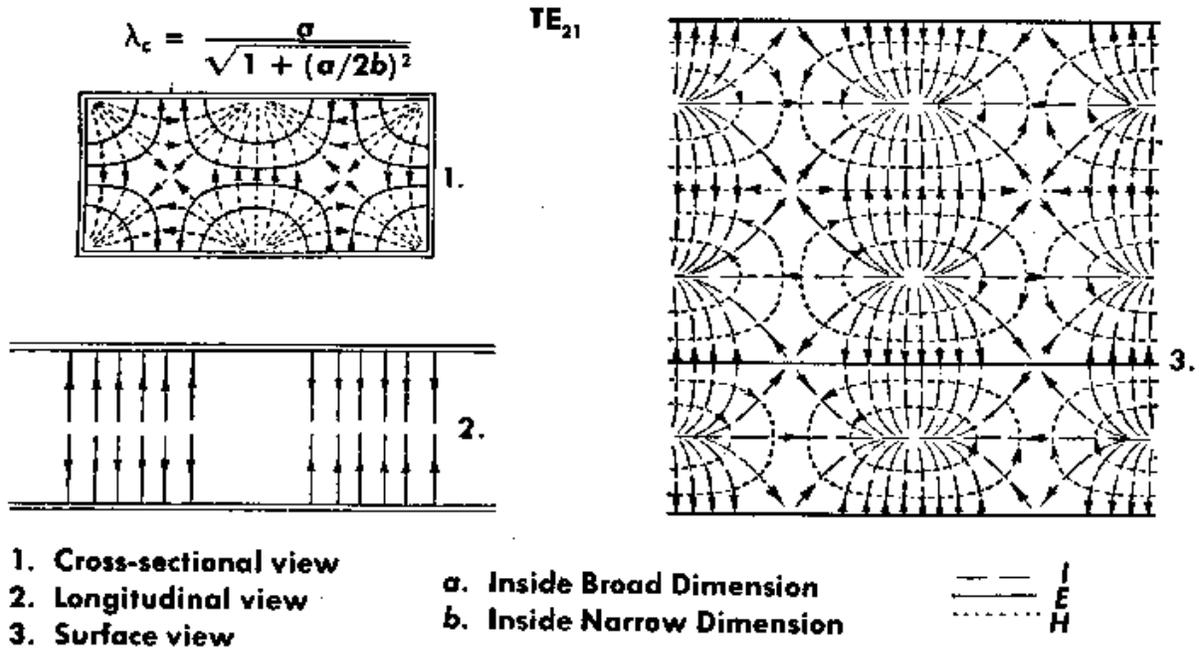


Figure 4 TE₂₁ Propagation Mode (Cut-off Frequency 59.6 GHz)

Measurement Test System Details

The test system was assembled to do measurements in WR-22, and contained the following parts:

- HP 8757A Scalar Network Analyzer
- HP 82025 Q Band WG Detectors (WR22)
- HP 8697A-H50 Sweeper 33-50 GHz

- Baytron WR22 20 dB Broad wall Coupler
- Various Q Band Loads, Attenuators
- Test System in WR22 WG Round Flanges
- Commercial Round to Rectangular Adaptors
- Tests Conducted
 - Return Loss & Insertion Loss (Through Attenuation)
 - 40 to 50 GHz

Some of the tests required the use of the “Bare” detector modules, which limits the Return Loss to approximately 20 dB. Other measurements allowed the insertion of a 10 dB waveguide attenuator and resulted in much improved Return Loss results.

Test Results

The first item tested was a 2 inch (5 cm) straight piece of WR-28 since this is the most basic requirement for connectivity. Both waveguide flanges are flat without any choke grooves and the results in Figure 5.

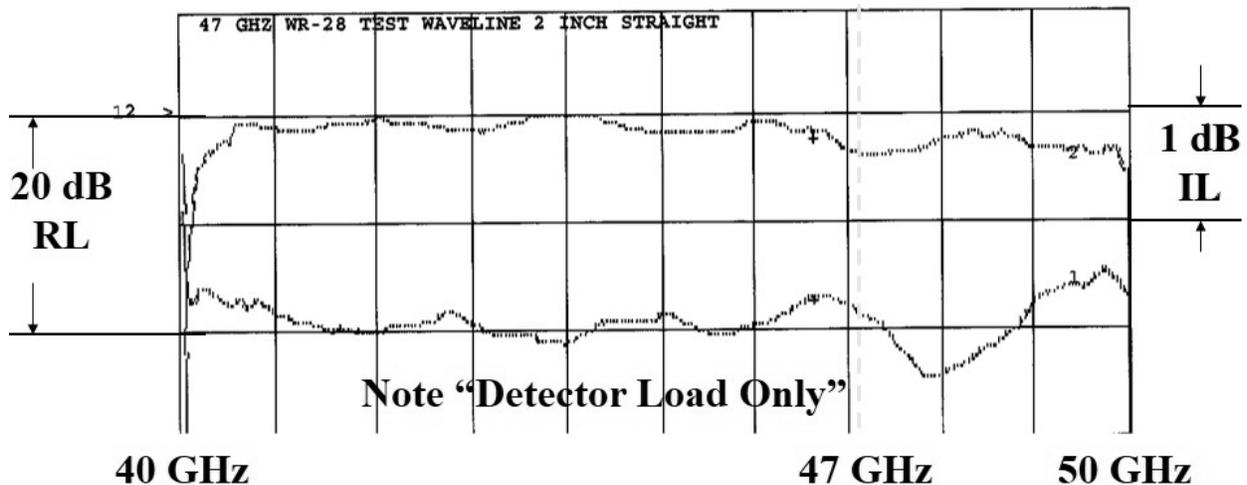


Figure 5 2 inch (5 cm) Straight WG Test Result

The next test was on a longer 9 inch (23 cm) straight waveguide as shown in Figure 6. Both of the straight sections show reasonable insertion loss and good return loss up to 47 GHz

Figure 7 shows the test result for a 4 inch (10 cm) corrugated “Flex” waveguide. It exhibits considerably higher ripple in the Insertion Loss than a straight waveguide but still has good Return Loss. The Insertion loss at 47 GHz is still reasonable, but we cannot be sure that will always be the case.

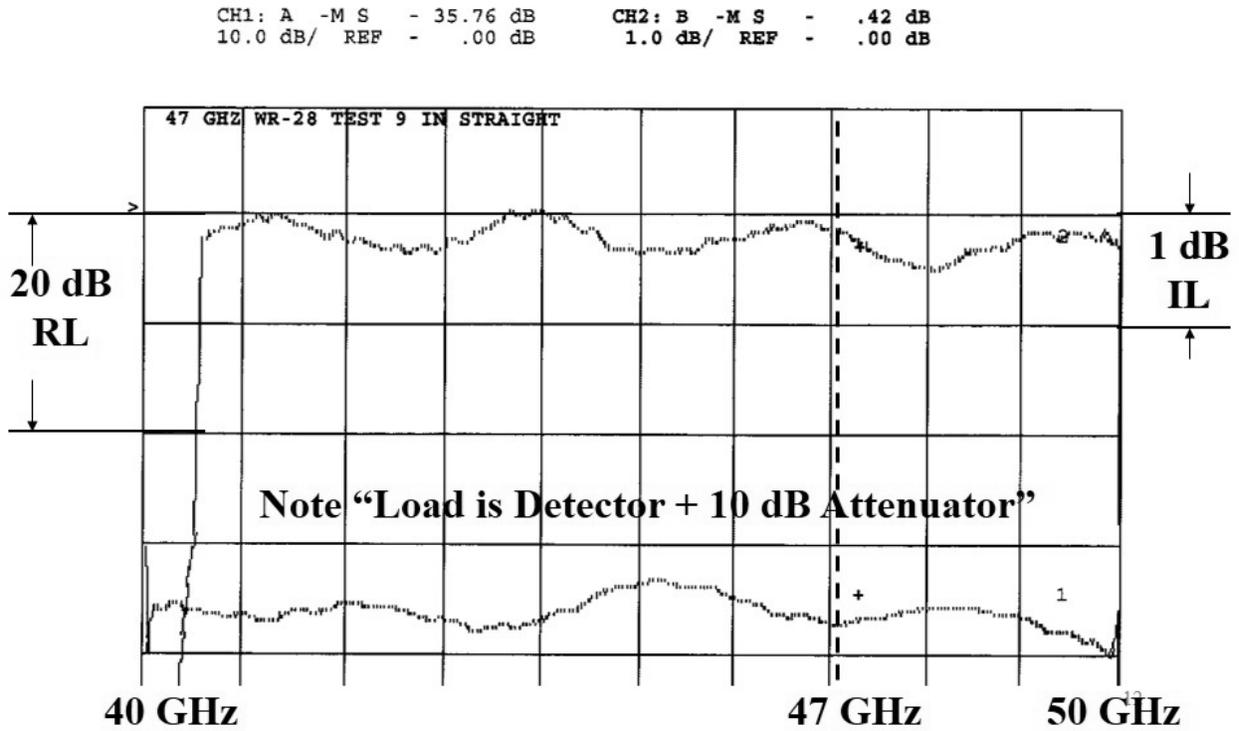


Figure 6 9 inch (23cm) Straight Waveguide

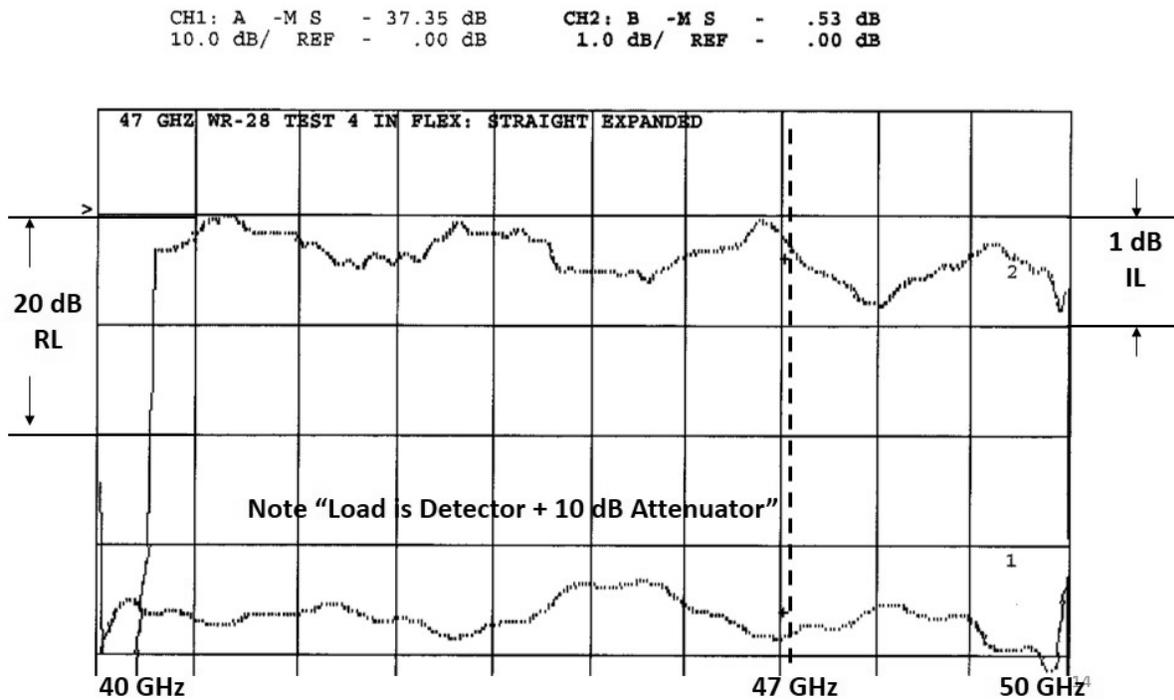


Figure 7 4 Inch (10 cm) Flex Waveguide

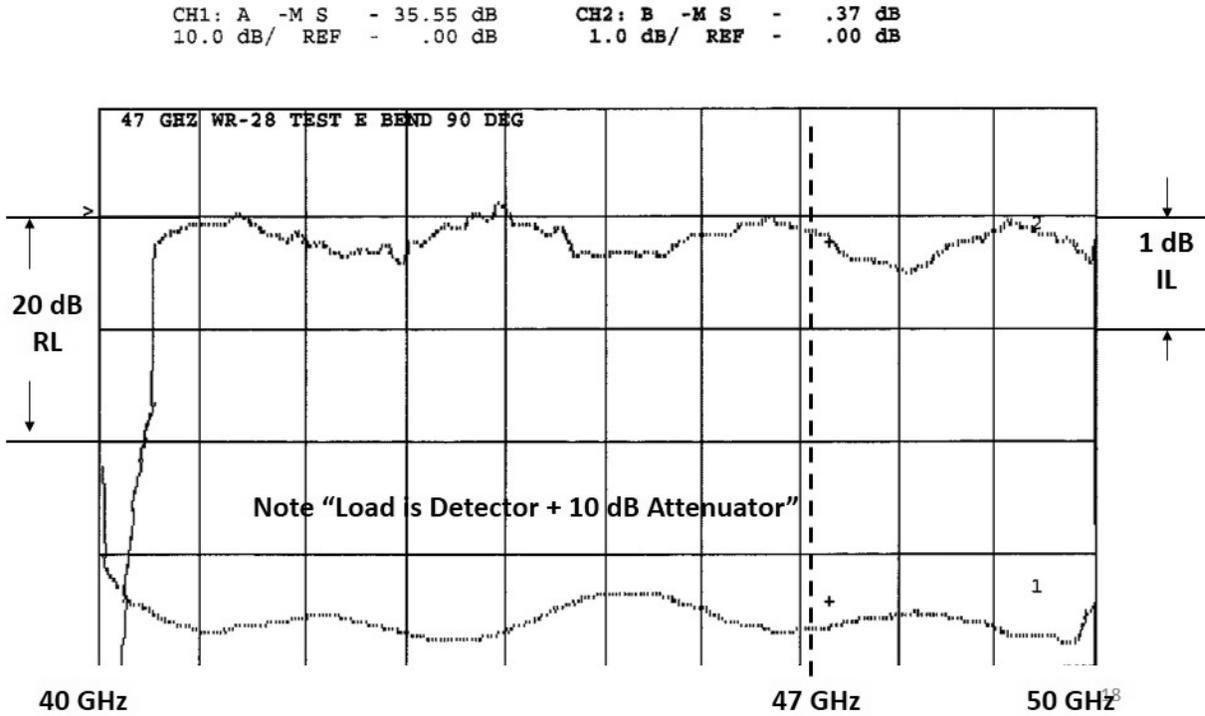


Figure 9 Rigid Waveguide E Plane Bend

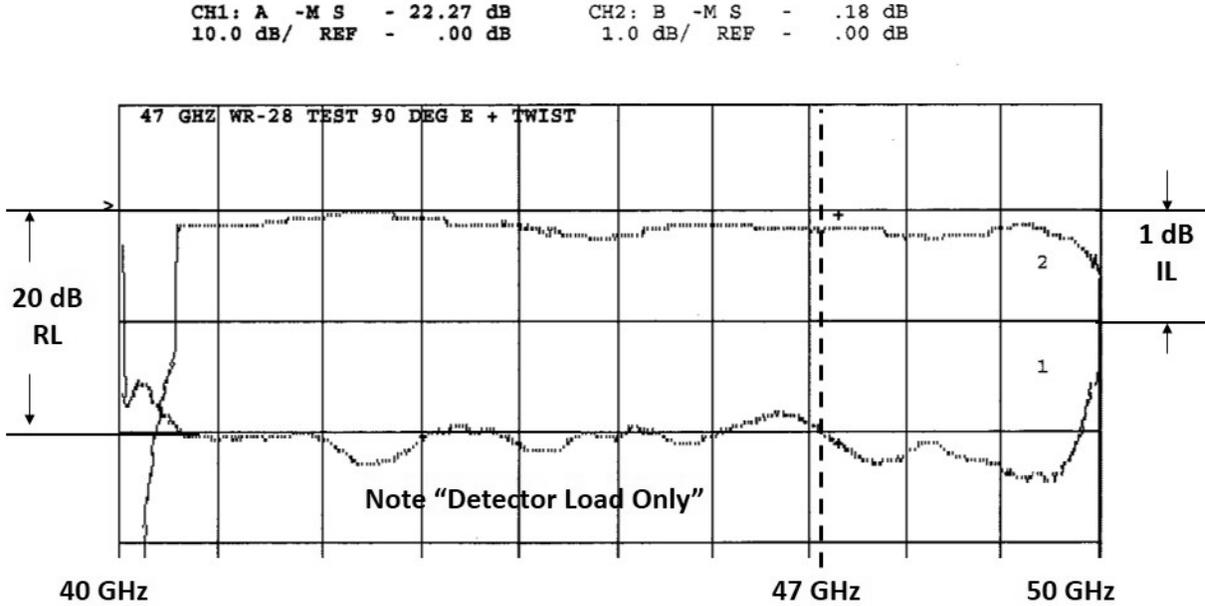


Figure 10 Rigid Waveguide E Plane Bend + Twist

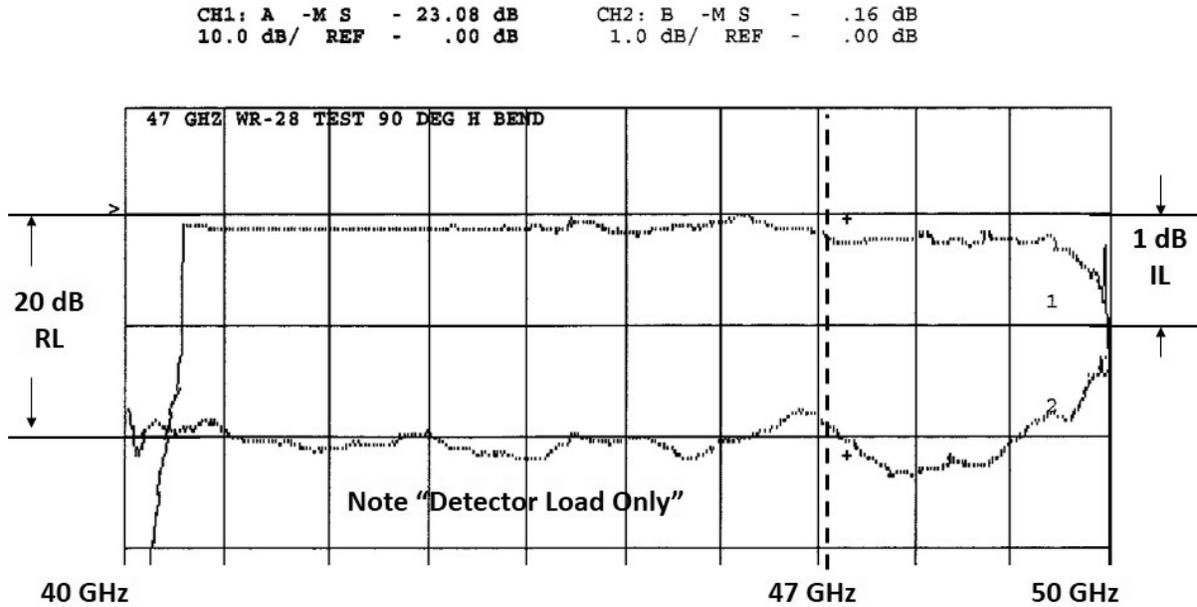
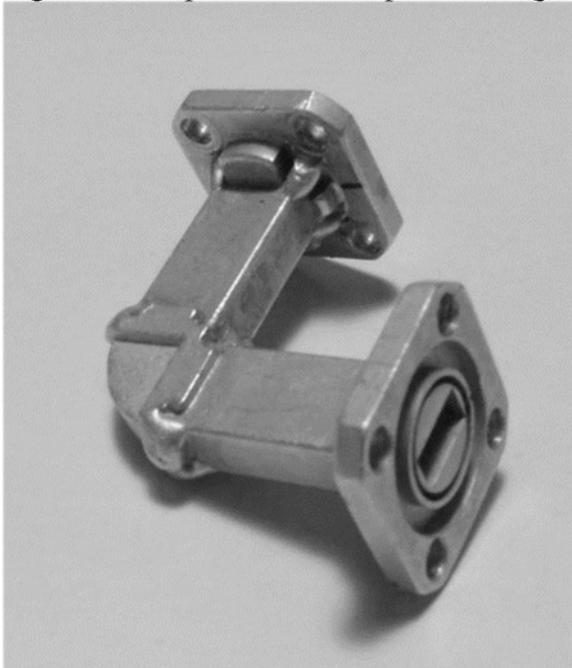


Figure 11 Rigid Waveguide 90 degree H Plane bend

Figure 12 is a picture of an H plane 90 degree bend with a "cast elbow" bend. This piece also has a choke flange which the previous 90 degree bends did not have. The response shown in figure 13.



Very clearly the Insertion Loss response falls off very quickly above 40 GHz with approximately 1 dB of loss at 47 GHz, and with a minimum at 48.5 GHz. Looking at the Return Loss curve at 48.5 GHz there is a very strong resonant looking response and then again near 50 GHz.

We were concerned with the possibility of moding with an H Plane bend and this would appear to be it. The previously shown H plane bends (Figures 8 & 11) were of swept radius (gentle bends). The cast bend is abrupt with a short radius.

Figure 12 Cast Elbow H plane bend with Choke

Flange

This is also the first response of a waveguide using a choke flange. Al W5LUA has suggested that the choke flanges do have an influence and perhaps a better response could be obtained by filling in the cavity of the choke in order to change its frequency response. More on this later.

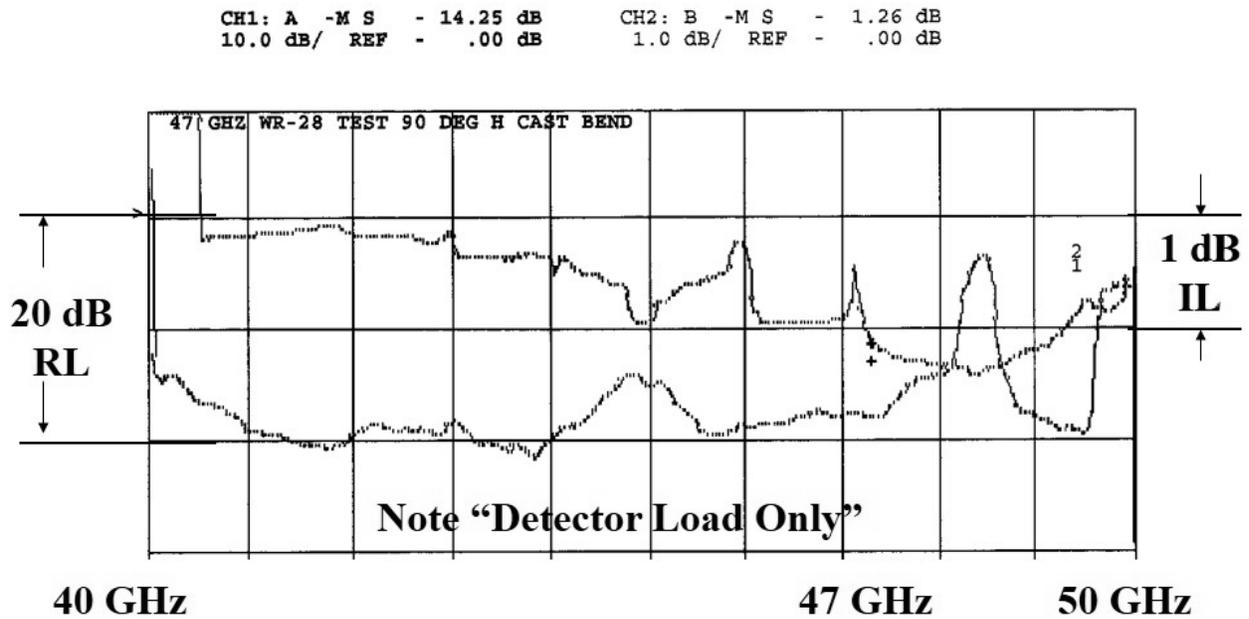


Figure 13 Cast Elbow H plane Bend

The next tests were of various waveguide loads. The first is a low power (1W) unit and its response is quite good at over 20 dB Return Loss as shown in Figure 14.

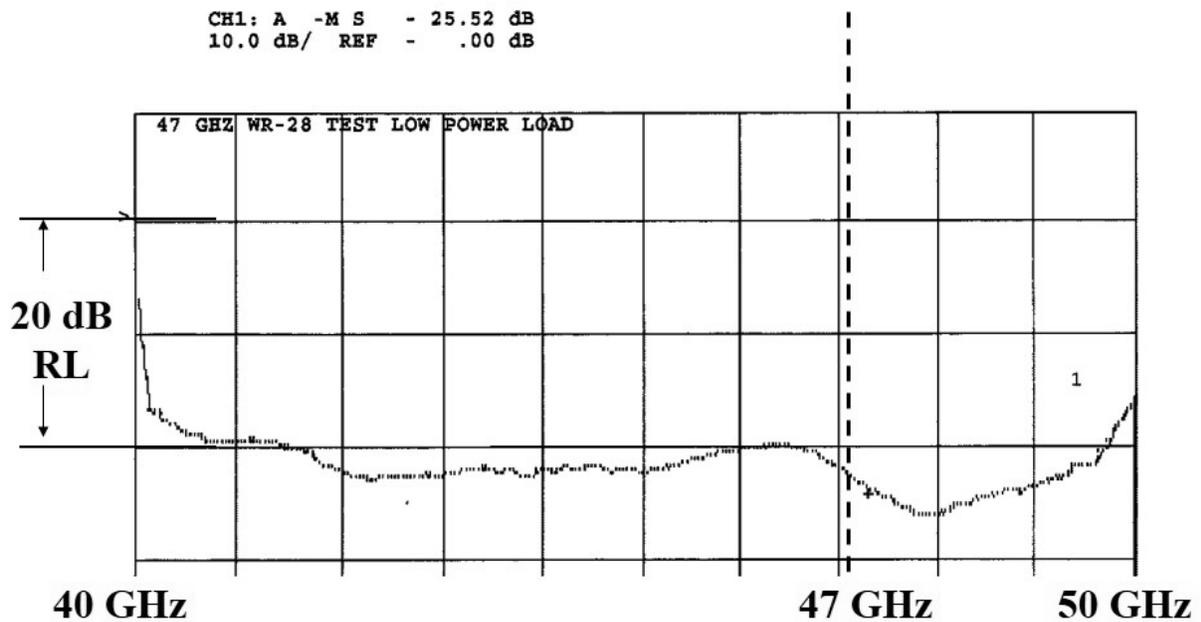


Figure 14 Low Power Load

The next load to be tested was a 50 Watt unit with large fins. Its response is shown in Figure 15.

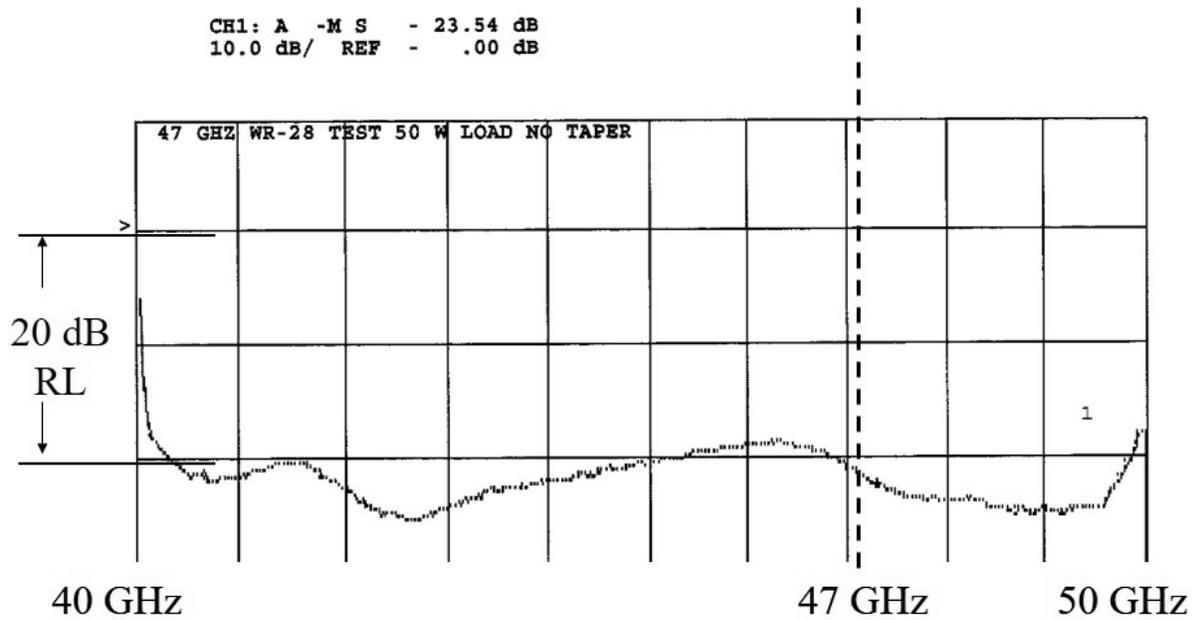


Figure 15 50 Watt Load

In an effort to improve the frequency response, a smooth taper transition from WR-28 to WR-22 was inserted and the result is shown in Figure 16. That is a very nice improvement from 22 to 26 dB and a much flatter response! Another alternative would be to use a $\frac{1}{4}$ wavelength thick metal step transformer as shown in Figure 17. The bandwidth will not be as good as with a taper, but sufficient for our use.

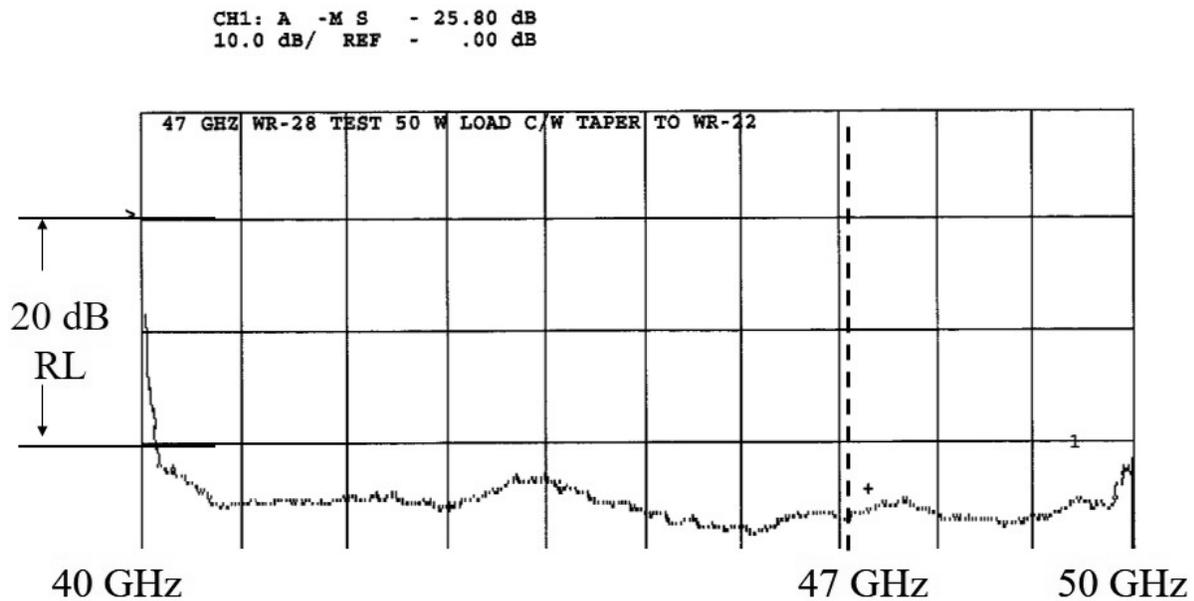


Figure 16 50 W Load with Taper Transition to WR-22

This switch shows approximately 0.5 dB Insertion Loss, not much different than a straight piece of WR-28 waveguide, but the Return Loss is only about 18 dB and has significant ripples vs frequency. Notably the response of Insertion Loss and Return Loss deteriorate significantly between 49 and 50 GHz. We had seen similar behavior with nearly all of the waveguide pieces tested previously! I had attempted to do some waveguide tuning to improve both responses at 47 GHz but was unsuccessful.

The next switch tested was a Waveline 1078-H Manual H plane unit with flat faced flanges and its response is shown in Figure 19. The response of the switch is quite remarkable with decent Insertion Loss of 0.4 dB and Return Loss of 18 dB, but the response deteriorates considerably at 47.5 GHz and then again at 49.5 GHz. It appears that internally the switch is probably has a fairly abrupt H plane bend.

The final waveguide switch was a Relcomm Tech E Plane unit shown in Figure 20, which appeared as a “New Old Stock” surplus from New Zealand. These units were small, powered from 8Vdc and had latching characteristics, which made it ideal for amateur portable operation. This switch does have choke flanges.

The response varies very dramatically with a serious deterioration peaking at about 47.8 GHz. The Insertion Loss is about 2.2 dB and Return Loss of 12 dB. This was very disappointing! I made no attempt to tune this out. Since it internally is an E Plane Switch it is surprising to see this response. As discussed earlier Al W5LUA had questioned the role the choke flanges has had in this poor response. It still is a terrific switch for 24 GHz!

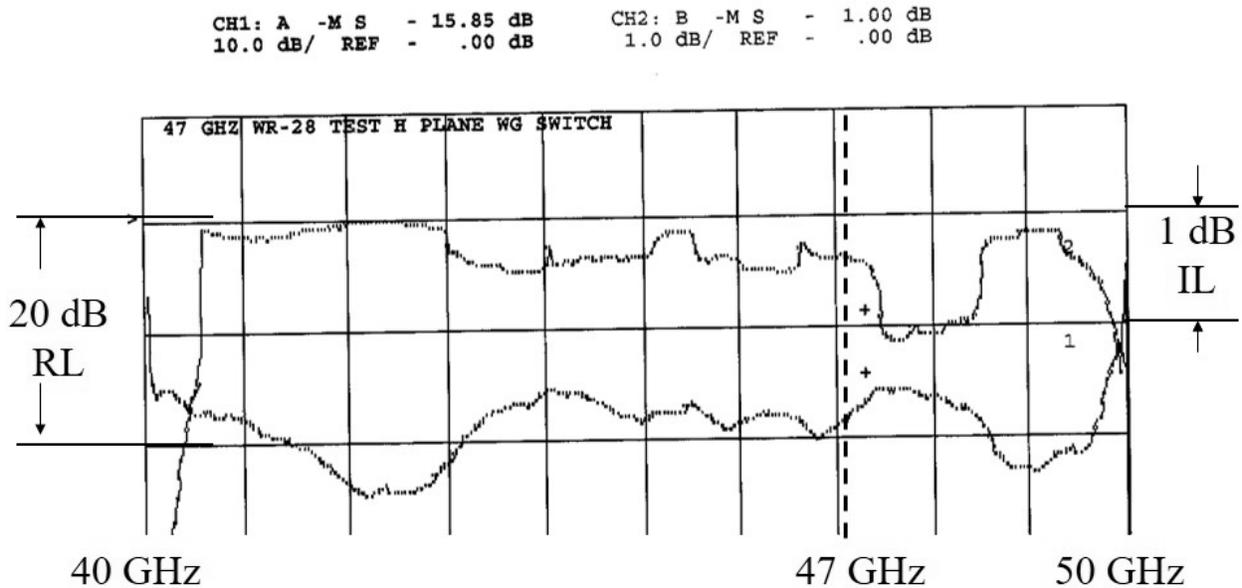


Figure 19 Waveline 1078-H H plane Waveguide Switch

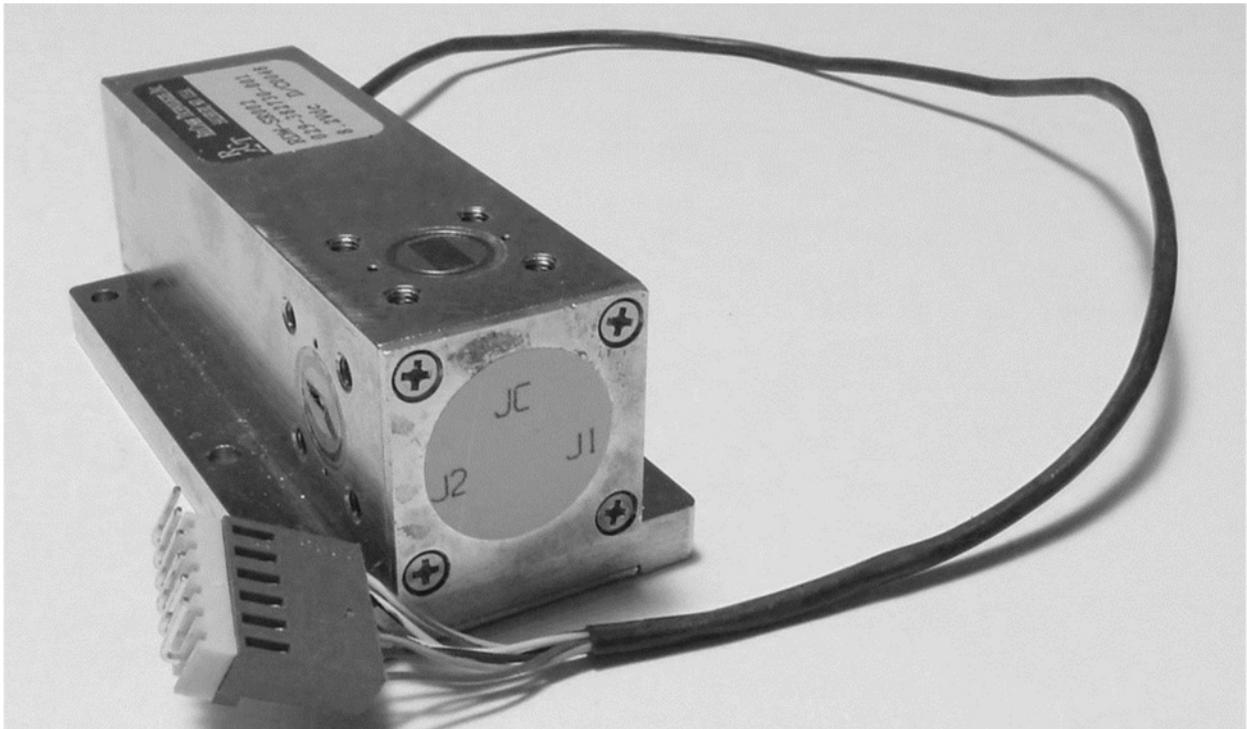


Figure 20 Relcomm Tech E Plane Waveguide Switch

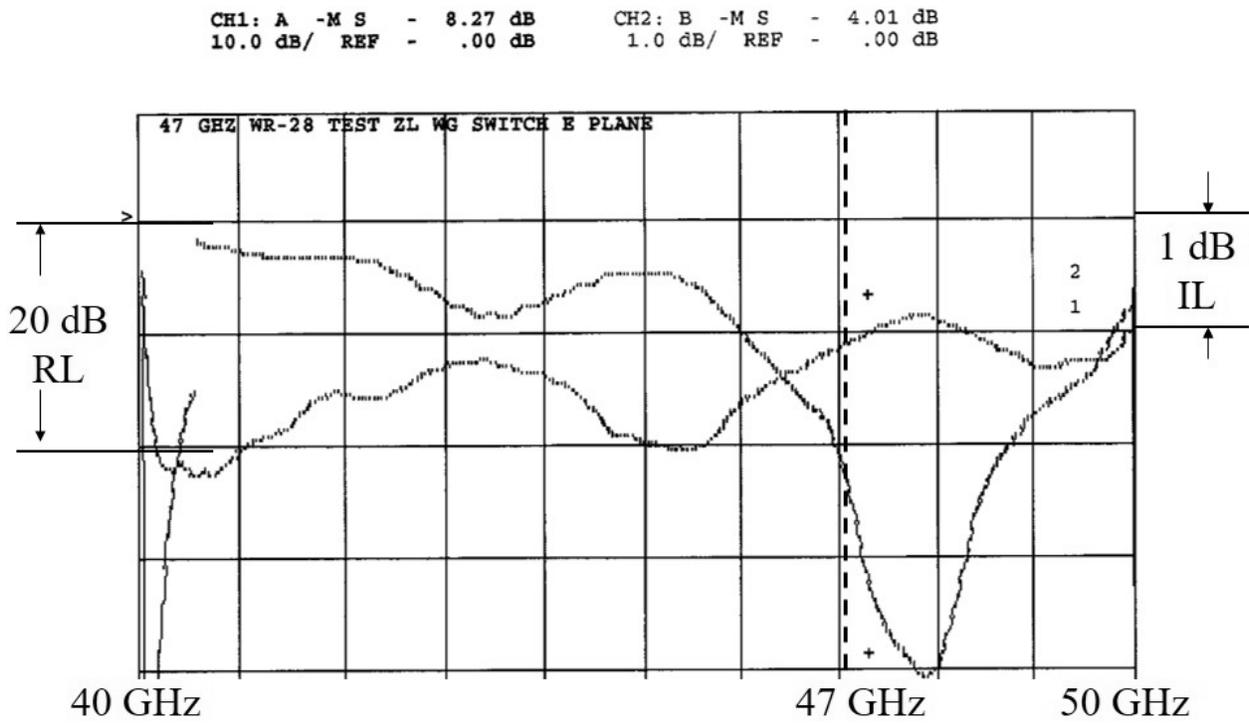


Figure 21 Relcomm Tech E Plane WG Switch

The final part that was tested was an MCS R382-B Broad wall Directional Coupler shown in Figure 22 I could not find a data sheet but from the part number I assumed it was a 6 dB coupler. The response is shown in Figure 23. The coupling factor looks as if it's about 12 dB instead of 6 dB, and at 47 GHz the coupling is 7 dB. The Return Loss of the transmission port is about 36 dB and does show a peak decline at just over 48 GHz. Through port Insertion Loss was not examined.



Figure 22 MCS 382B Broad Wall Directional Coupler

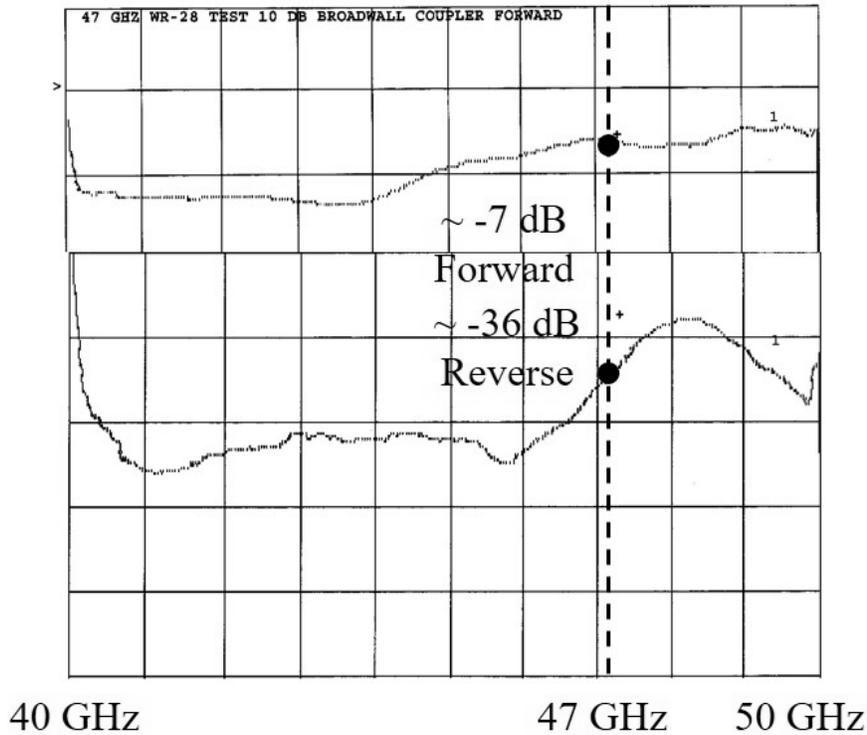


Figure 23 MCS 382B Directional Coupler

47 GHz WR28 Waveguide Summary and Recommendations

- OK to use WR28 If the lowest loss is not required
- Use only short and straight sections If possible
- “E” plane bends are best, and large radius bends are preferred
- “H” plane bends may be OK....but test them!
- “Cast” 90 Degree H plane bends are very bad. E Plane bends may be OK, but were not tested.
- High power loads are OK, but are better with a taper or step transition to WR-22
- Waveline switches are generally OK, with E plane types being the best. The Relcomm Tech waveguide switch was very poor.
- Waveguide components with Choke flanges are suspect. Test them!

Possible Follow-up Work

- Perform similar Insertion Loss and Return Loss tests with interface to a WR-19 test system

Using Ku band PLL LNB's

Doug Millar K6JEY

drzarkof56@yahoo.com

In years past Ku band LNB's were used on 10GHz as amplifiers. The LNB's were designed as superhetrodyne front ends for satellite receivers. They had several stages of RF amplification at a good noise figure but had dielectric resonator oscillators (DRO) for their LO's. DRO's are excellent oscillators but are relatively unstable. The latter made them unusable as 10GHz front ends for us, although plenty were converted to RF amplifiers.

In more recent times there have been two improvements in LNB's. The first is the advent of "Universal" LNB's that cover the upper half of 10GHz. The second is that chips have been developed that include the mixer, PLL and IF amplifier. A 25MHz crystal locks the PLL and provides usable stability. The modification to use a TCXO makes the modern LNB's an excellent front end, both as a receiver and for a spectrum analyzer.

As I was working on the latter idea, it occurred to me that the LNB's would make an inexpensive 10GHz front end for a lower frequency spectrum analyzer. That arrangement might allow microwavers to spend more money on a quality "back end" and use an LNB for the front end, instead of buying a second fairly poor microwave spectrum analyzer. Along the way several other ideas crept in such as a GPS source locking module, modifications for an external input and changing the IF frequency to 432 MHz

The use and modification of several LNB's

A universal LNB is the easiest to use and translates 10.368GHz to 618MHz as an IF. The LNB usually comes with a 25 or 27MHz crystal for its PLL. By changing to a crystal of a different frequency an IF of 432MHz can be obtained. Straight out of the box with a 618MHz IF the LNB can easily receive local beacons, etc. With the addition of a dish, some have even copied the DL0SHF beacon.

The first LNB is the Norsat HS1057HCN . The model number says the story. It is stable to 5 kHz, has a .7db noise figure, is tuned for the C, or lowest band, and has an N connector on the output. There are lots of different types and the buyer should be careful. This particular model has a WR75 input. It is designed in the 1990's style of circuit and engineered quite well. The stability and repeatability are excellent with almost no drift observed. However at the low IF of 368 MHz, the sensitivity is down about 10db from the other LNB's. I have been using it as a front end for a low frequency spectrum analyzer and it works very well.

HOW TO ORDER

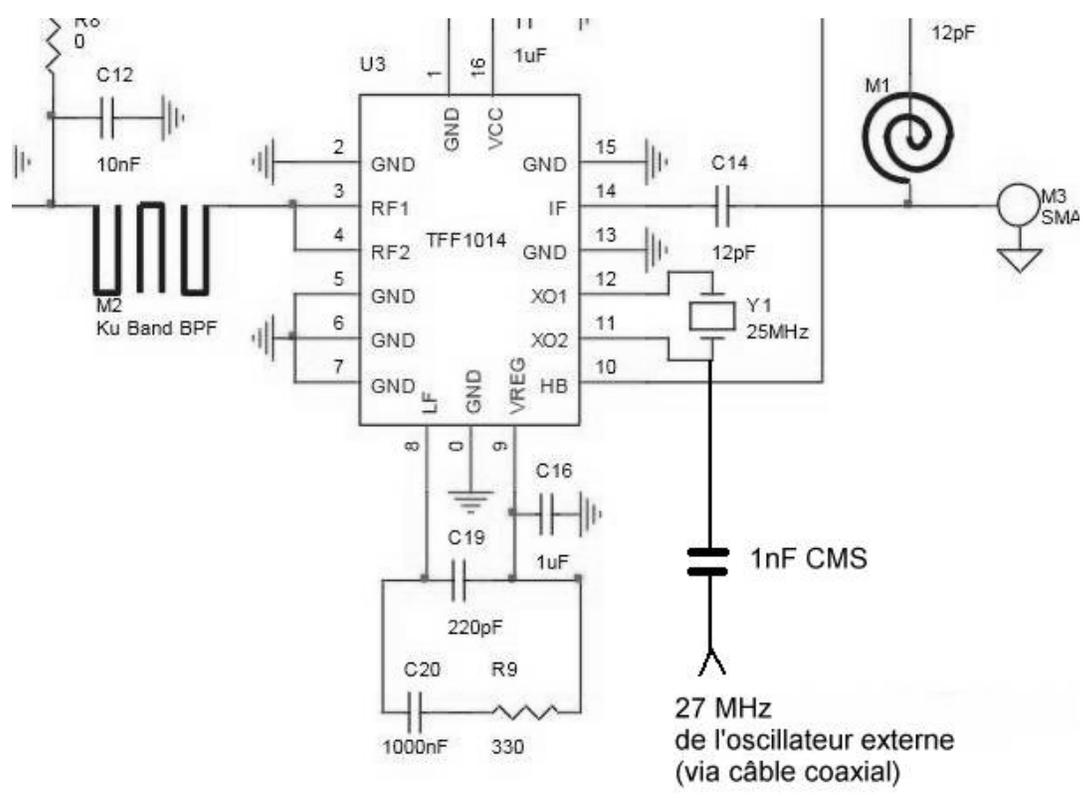
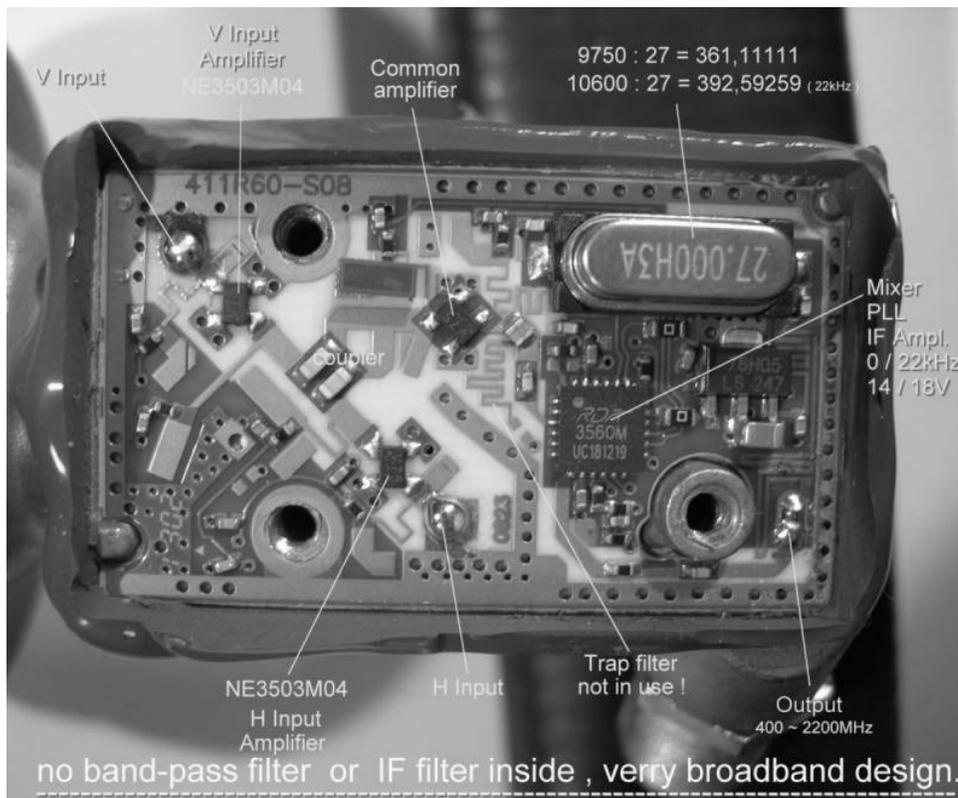
HS1048AF

CONNECTOR	F - 75 Ohm N - 50 Ohm
FREQUENCY	A - 11.70 - 12.20 GHz B - 12.25 - 12.75 GHz C - 10.95 - 11.70 GHz
NOISE FIGURE	9 - 0.9 dB 8 - 0.8 dB 7 - 0.7 dB
L.O. STABILITY	5 - ± 5 kHz 4 - ± 4 kHz 3 - ± 3 kHz 2 - ± 2 kHz
LNB SERIES #	



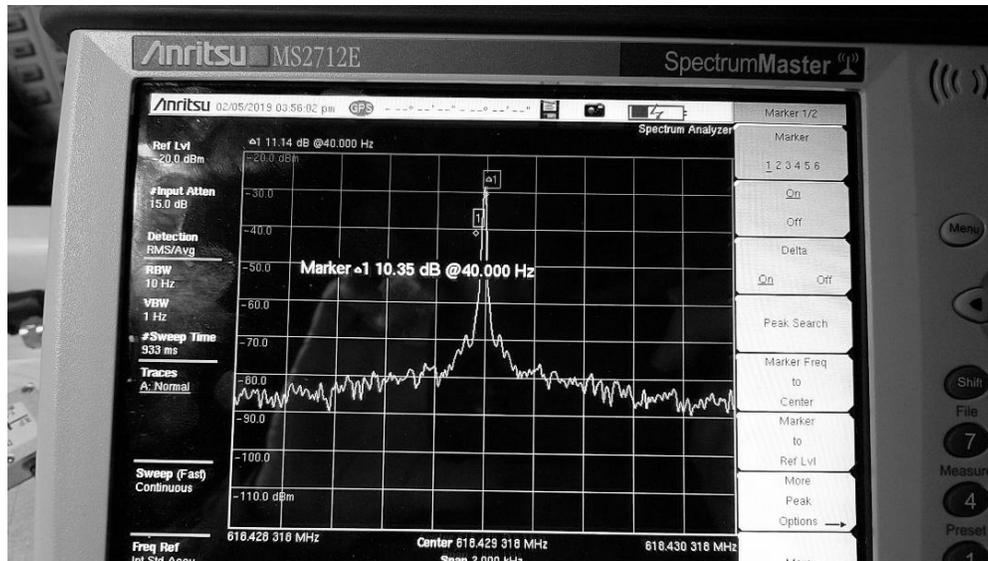
Amiko L-104

The L104 is a better candidate. You can see from both of the pictures that these are very simple- RF amplifiers for each polarization are combined into a second RF stage. An interdigital filter follows and goes into the PLL/mixer chip. There is no output filtering. As can be seen from the pictures, modification is pretty straightforward. However, soldering is a different story. I ended up using a digital microscope soldering station and the finest wire I could find. It saved me making a mess of the LNB. With the modification the LNB drifts at 50Hz/minute – excellent stability.

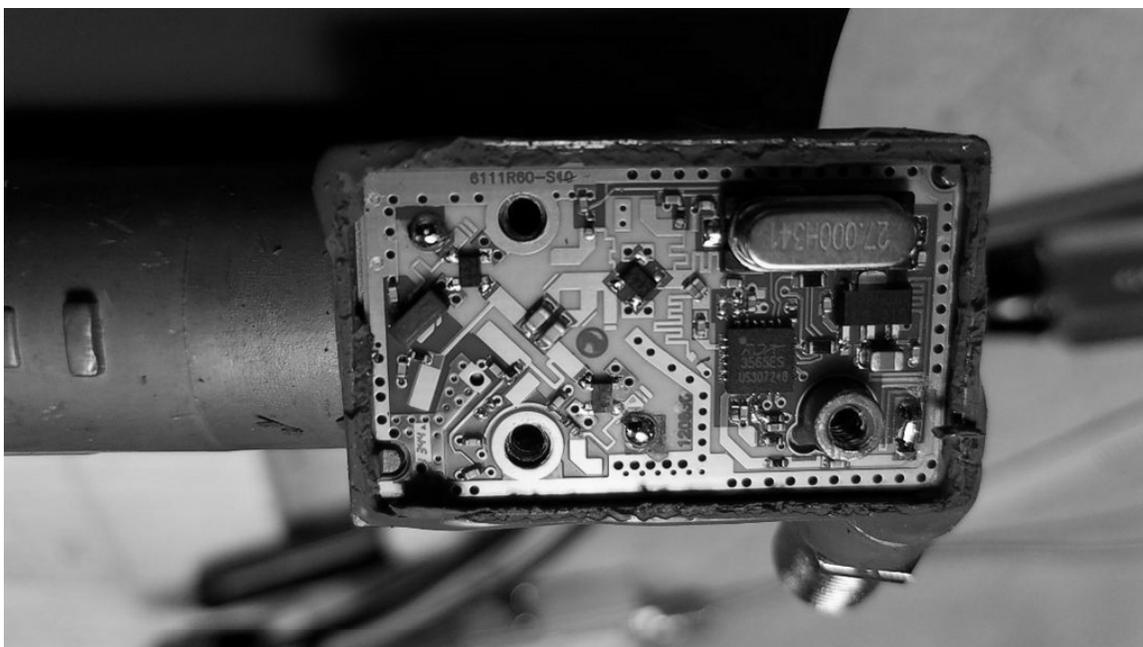


The TCXO is a good choice to stabilize the LNB as it is compact and does a great job. However, you are left with a 618MHz IF. By using the GPS disciplined oscillator by SDR Kits for \$100 you can generate any frequency between 400Hz and 800MHz. A frequency of 27.515MHz yields an IF of 432MHz (factor 361.11) for the L104 LNB. That should make the LNB a lot more useful. 1296 and 144MHz IF's are out of lock or sensitivity range. (From G4INT- Using an external LO Source with the Octagon PLL Low Noise Block 2016)

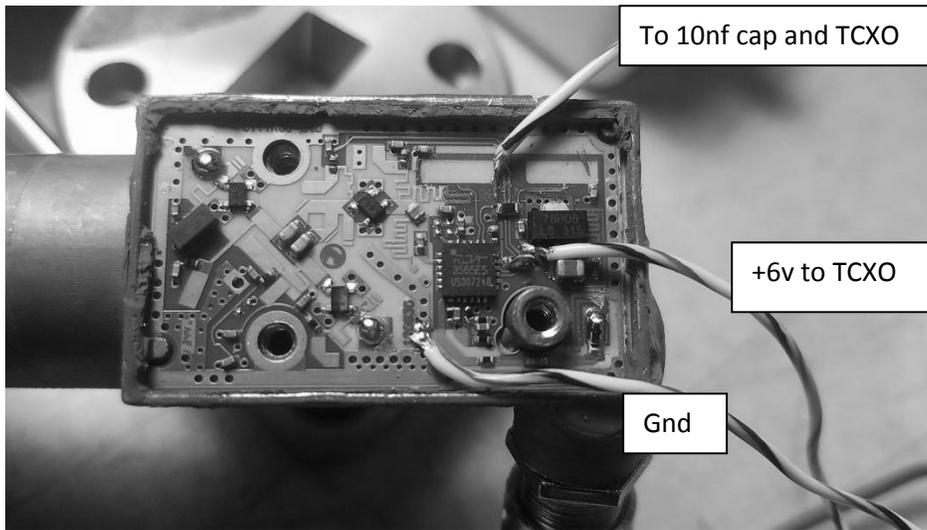
With a standard crystal the L104 LNB will drift 500cy per minute after a 15minute warm up. With a TCXO the drift is 50cy per minute after a 5minute warm up.



Anritsu analyzer at 618MHz IF showing a 40Hz drift during one minute.



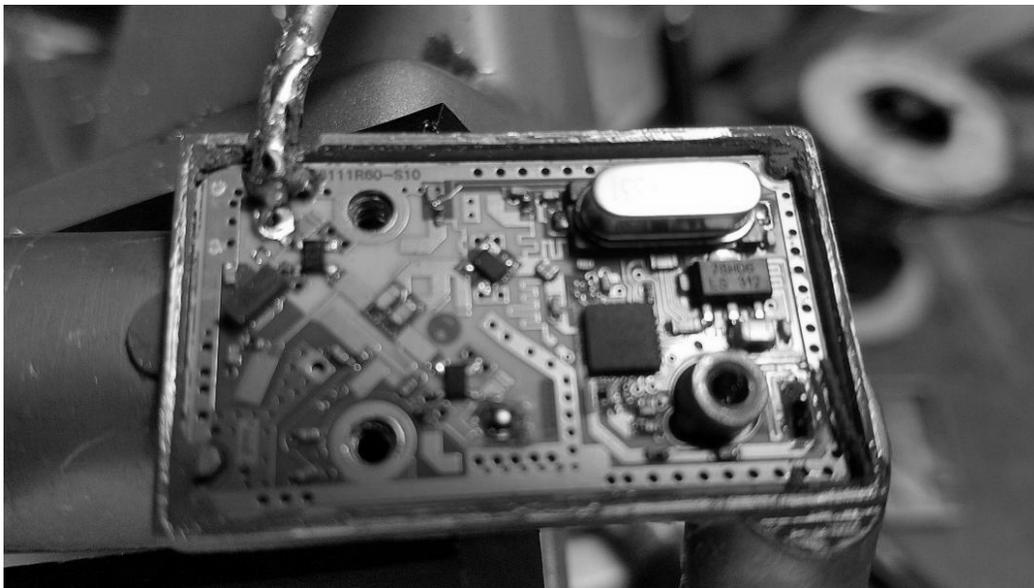
You can clearly see the modification. TCXO input, 6v from the regulator and a ground contact. Just use very small wire and RTV them down so they don't move.



External Input Modifications

One of the frustrations with working on the LNB's is that the input is a horn. For many applications, a waveguide or SMA connector would be more desirable.

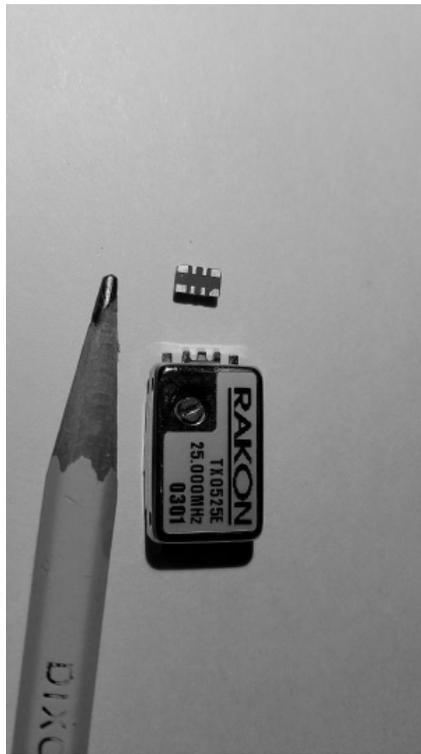
Here are a couple of examples of what can be done.



- Use the vertical input as the horiz. input is disabled by default
- Push through Vertical probe element

- File a hole in the seam of the case to pass the coax.
- Connect small dia. semi-rigid coax to V side
- For external connection
 - Use coupling cap from the gate to the coax
 - Use an external DC block. (0.6 db loss)
 - Use an SMA to WR90 adapter.





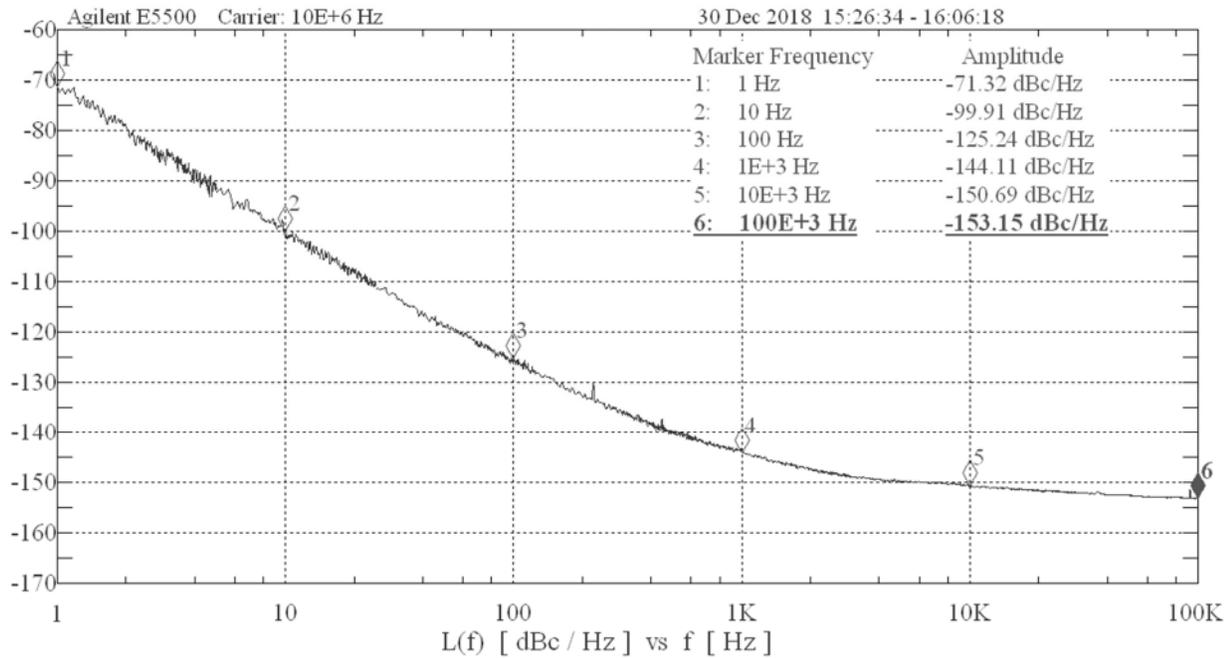
I recommend the RAKON TCXO as the other one is too small to be installed (by me).

SDR Kits GPS Ref. Source

One can use a GPS source from SDR Kits. It is truly a game changer. The box is small and can be run off of a phone battery. After it locks up its TCXO to GPS the antenna can be removed and the unit will go into holdover mode. The unit then becomes a standalone, very accurate source for any frequency between 400Hz and 810MHz. The phase noise is superb at -140db at 1 KHz. It should eliminate any drift problems in an LNB. Just couple it into the LNB through a 10nf capacitor.



Leo Bodnar Electronics mini GPS clock phase noise at 10MHz



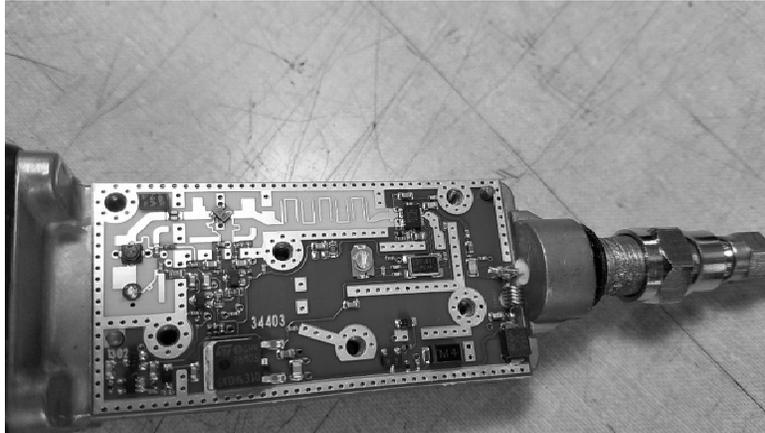
Other LNB's

Jan, OK2ZAW offers a complete package with an LNB, TCXO and outboard Bias tee. The Tee has an attenuator, fuse and pilot light. All for \$55. <http://ok2zaw.blogspot.com/search/label/10GHz> The output is on 432MHz. In general most LNB's can be changed to this Band and get good results. The 2meter and 23cm bands seem to be outside the range of the LNB's and don't do well.



N1320

This is about a \$20 purchase off of the internet. It has an internal filter which reduces the gain at the 368MHz IF and an uncompensated crystal at an odd frequency internally. At first glance it seemed a poor choice. However when paired with the RF Explorer Spectrum analyzer, it gains a new life. The analyzer has a minimum selectivity of 2.5 KHz and the 1320's crystal doesn't seem to wander off more than that. What it does is provide a compact and hand held 10GHz spectrum analyzer with a WR75 input. Probably the best deal in the presentation.



Applications- Receiver

Alinco DJ-X11

The first application is with a broadband receiver. I had an Alinco DJX11 hand held receiver which tuned 618MHz easily. In comparison it was much more stable and received better than the Kenwood TH F6 handy talkie. With a bias tee for the LNB it is still my favorite set up. I'm listening to the local beacon 50mi away.



SDR Dongle

The second application is with an SRD dongle. You can use whichever dongle works at 618mzh and it will work fine. Mind the input level or overload, though. The disadvantage is that you need a laptop to make it work. An Android or tablet application may make it more portable. With \$10 spent for the LNB and \$20 or so on the dongle setup ends up being inexpensive. The RSP 2 is an excellent full featured dongle and so is the Fun Cube Pro+.



From Graham G8HAJ, using a dongle, bias tee, and L104 LNB.

Receiver considerations-

Recently I did a comparison between the Alinco receiver and my old FT 817 transceiver as a back end receiver for the LNB's I used OK2ZAW's unit on 432MHz and compared the Alinco and Yaesu receivers. The Alinco worked OK but the FT817 was much better- stronger, more sensitive and tunes easily. The 817 sounds like an obvious choice, but I think it depends on your application. I suggest that if you want to do monitoring, use the dongle approach, or a Flex 1500. IF you want an extra receiver to take to the mountaintop or for testing in the garage, the Alinco is very handy and the battery lasts for days. The 817 is better, but may be better employed elsewhere. In addition, transmitting into the back end of an LNB is destructive for both the LNB and the 817.

Applications- Spectrum Analyzer

A 10GHz spectrum analyzer can be very expensive or very old, or both or big and heavy. Of course you may luck into something wonderful. Great, I hope you do. I have fond memories of my years with an HP 8551, Polarad SA 84W (just kidding on that one), various Systron Donner products, several AIL's, Tek 494AP, Anritsu and others. One could get one of those because of nostalgia or limited funds, or spend a bundle for a nice nearly new one, and I can't really say much against any of those choices (except for the SA84W). However, the LNB offers the use of a highly portable and more useful analyzer like the ones described below on 10GHz.

It seems to be productive to invest in a very good lower frequency analyzer and add a microwave front end to it. In this case I chose an Anritsu MS2712E. It is a current model that is available surplus for around \$1500. It includes GPSDO frequency setting, USB and Ethernet output/input, and a very large screen. It is also battery operated and portable I also recommend the Anritsu MT8212B. They often come with excellent accessories. Don't get the A model. There are there models and manufactures that are cheaper and can also work.

The Norsat Front 10GHz front end

The first application uses the Norsat in an open chassis with a variable attenuator. The bias tee is hidden next to the Norsat. The rig works well and gives about a 200MHz wide spectrum. I used two fixed attenuators and a variable attenuator to equalize the gain between the input and output. I put a chart on the front to note settings for reducing the input by 0-40db.



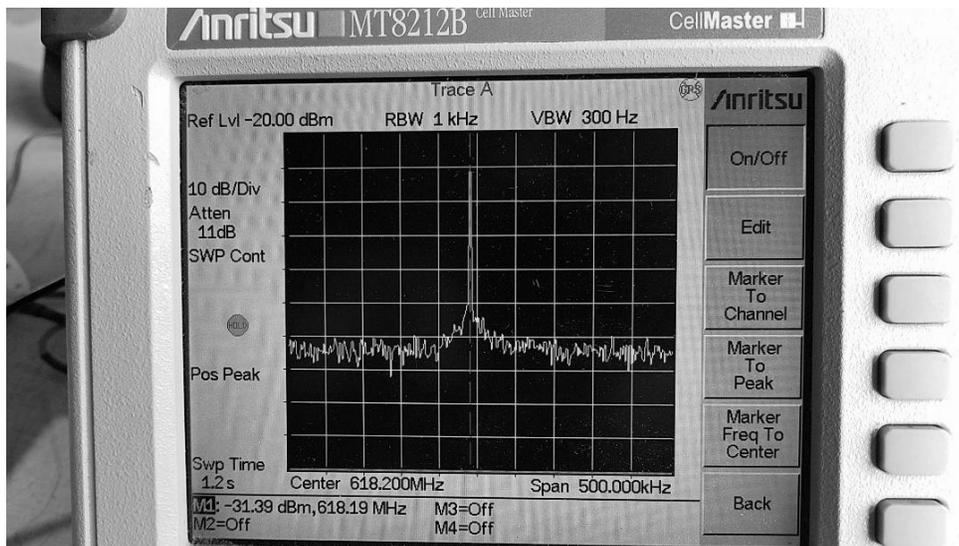


The Norsat and the RF Explorer

This begins to be a real hand held 10GHz analyzer.



Here are several more pictures of analyzer applications.



Anritsu MT8212B

The above analyzer is probably the best deal in town. For less than \$1000, you can get a great spectrum analyzer, vector network analyzer, power meter, gps, and more.



HP 8563A

Here is the analyzer adapter on an older HP 85 series analyzer.

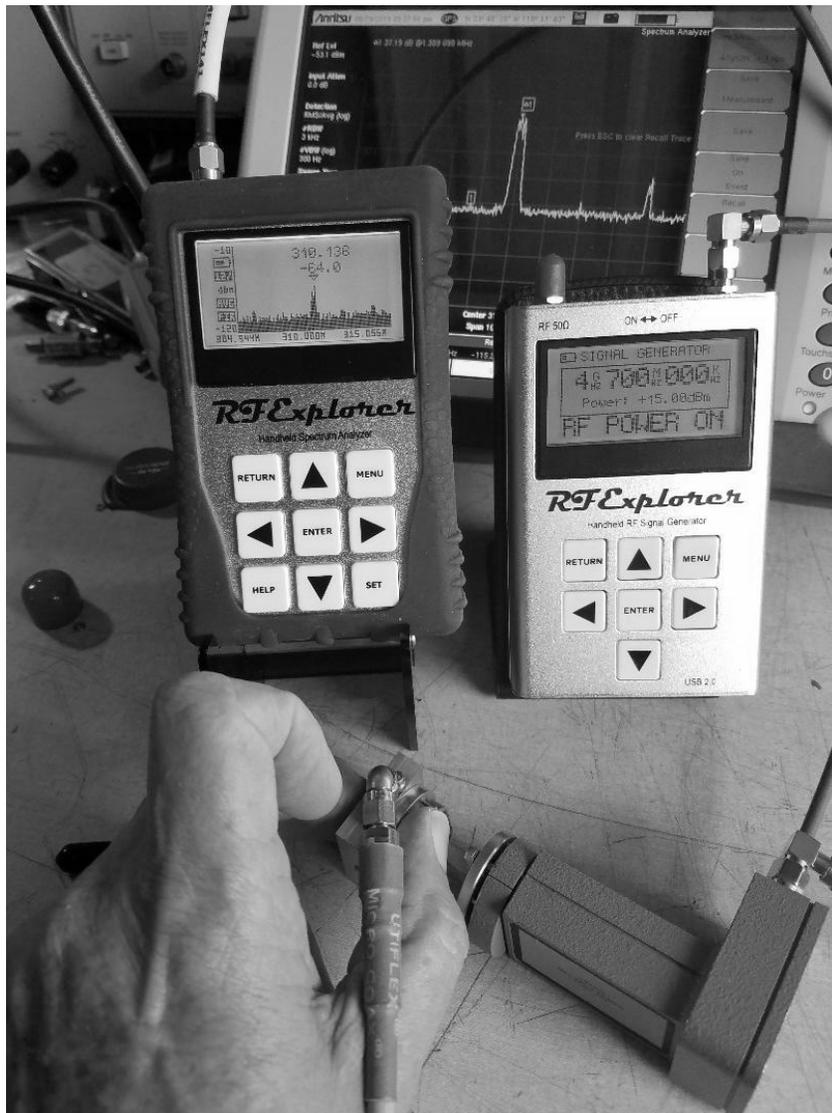
Looking at even Higher Bands- 24 and 47GHz

Having taken things this far, I wondered if I could make the idea work on the higher bands. I took an HP11870k external mixer (18-26GHz) and measured what the HP analyzer fed it. It turns out that on 24GHz, 8mw at 4GHz gets the mixer going and gives a good signal on both the RF Explorer and the Anritsu. I use an RF Explorer signal generator to supply the LO. Be advised that RF Explorer has a modification board that can be added to the generator that improves the output. Mine has the board in it. Here is a hand held 24GHz spectrum analyzer for about \$500.



47GHz

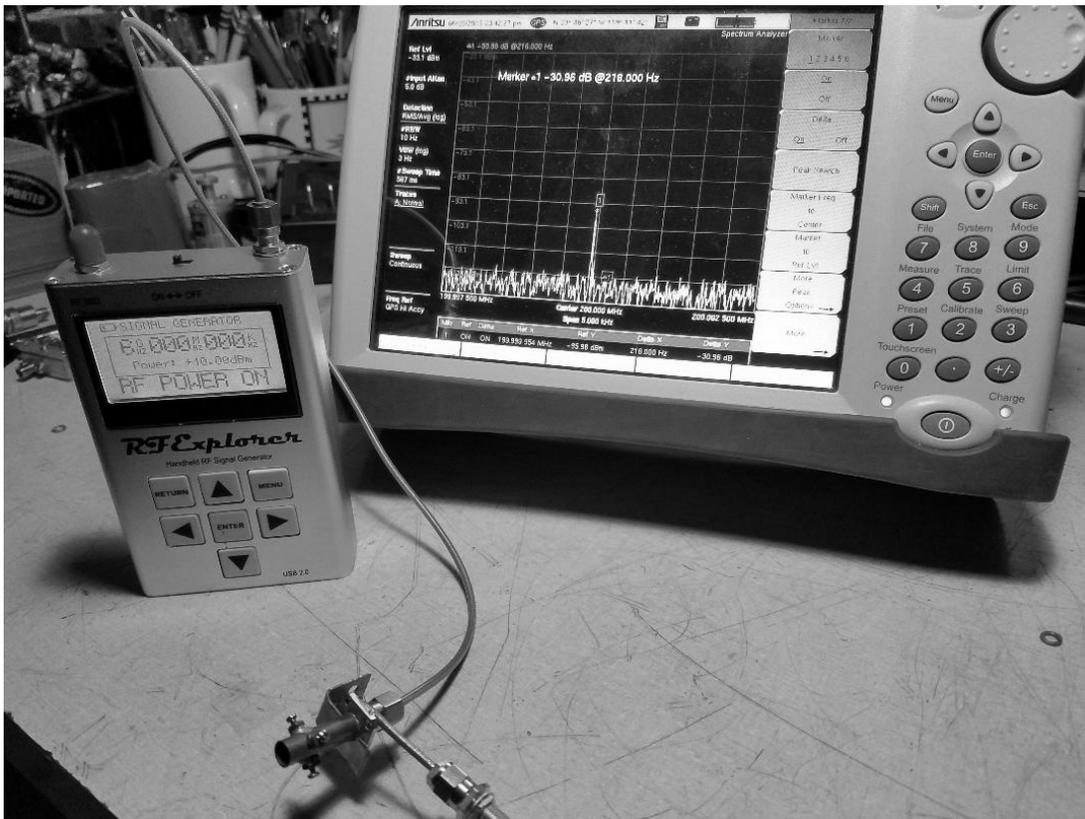
The next step was to see if I could get 47GHz working. The HP puts out about 40mw at 4.731GHz. I was able to get 36mw out of the generator and tune the analyzer to 310MHz. I used an HP8673B signal generator with a Spacek RF doubler on the output. The Mixer is an HP11970Q for 33-50GHz. The picture also shows a saved image of the setup on the Anritsu. A simple, hand held, battery operated spectrum analyzer for 47GHz.



Cheaper Mixer for 24-79GHz

If the cost of the mixers is a little steep for you there is an alternative. A discussion of mixer performance and the instructions for this mixer are in the reference page.

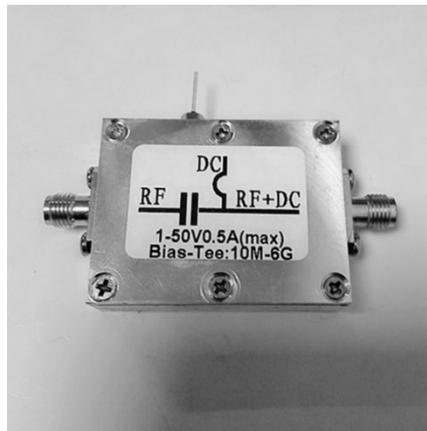
Kerry Banke K6IZW used some hobby brass and a Qualcomm mixer to fabricate a simple three port mixer for the upper bands including 79GHz. The cost is probably \$10. The results speak for themselves. 24GHz RF input with a 6GHz LO input at 8mw. The mixer also works well at 47GHz, but not tested at 79GHz.



There are obviously some compromises involved in using the mixer scheme above. I think the limitations are outweighed by their economy, versatility and reliability. All of us do our building in individual ways. That is part of the beauty and challenge of microwaves. I present the above and wait to see what you can do with it.

Extras-

Bias tee- This is needed to feed power into the LNB. You can use almost anything, but the pictured bias tee is available on Ebay or \$15 and contains an amazingly complex circuit. I would suggest that you find a source for 0.1uh RF chokes in case you short the B+ to the LNB. It is easy to do with F connectors. A resettable fuse is also good.



Portable Dish-As of this writing, this dish is available. It has 27db gain. Of course any other dish can be used. This will provide a big improvement in gain and distance, but takes some getting used to. For \$15 it is worth it.



Conclusion-

For about \$50 you can put together a respectable 10GHz receiver and enjoy activity on that band. You can also improve the frequency range of your low frequency spectrum analyzer for a small investment. If you want to improve on the basics, options are reasonably priced and make a big difference.

Since beginning this project about a year ago with the above as a goal, the project has spread out into some interesting areas- external sources, external connectors, hand held analyzer ideas and options for the higher bands. I hope these are helpful, and I hope they inspire you to do some exploring. If you come up with more ideas or better engineering and soldering, I'd appreciate hearing about your progress.

Happy microwaving and enjoy, Doug K6JEY

References-

Homemade mixer from Kerry K6IZW

<http://www.ham-radio.com/sbms/sd/47ghzmxr2.pdf>

K6JEY Talk on Mixer evaluation.

http://www.nitehawk.com/k6jey/harmonic_mixationers.pptx

RAMBO ROVING FOR MICROWAVES

Ben Lowe, K4QF

OCTOBER 3-5, 2019

I. INTRODUCTION

Many operators would like to set up stations for roving during microwave contest, but finding the best system components present a challenge. Several objectives are present for the system design, so this paper presents one way to configure a roving system. In configuring a system, the goals are to find a rugged container that will protect the system components from the weather and other elements in the environment such as heat and insects, keep the costs reasonable, and have a station that is easily transportable and versatile. By versatile, it is desirable to have a systems that is usable both roving and for other uses such as fixed station operation for both point to point communications and modes such as moonbounce (EME). The example shown here is for 1296 MHz, but the techniques are applicable to the other microwave bands as well.

II. THE SYSTEM CONFIGURATION

As is always the case, the initial thought is to run lots of power. However, if roving, the size and weight of the battery for operating a station with several hundreds of watts quickly modifies that thought. Especially when microwave roving, in some situations the operator may be faced with hauling equipment up a tower such as some of the abandoned fire tower, a lookout tower on a mountain top, or accessible by hiking $\frac{1}{4}$ to $\frac{1}{2}$ mile to get to the edge of an elevated location that is beyond the absorbing foliage which can present severe signal attenuation. Placing the equipment on the ground and running feedline up a tower necessitates using heaving runs of low loss coax and a rotor, both which complicate the rover operation particularly for one person when out roving. Generally, the microwave rover will mount the station on a light weight stand such as an aluminum speak stand, run lower power in the order of 20 to 25 watts, and use a smaller battery as a power source for the station. This configuration permits the rover to hand rotate the antenna, use minimum length transmission and power lines, and listen to the I.F. radio while peaking the antenna on the remote station sometimes in both azimuth and elevation.

With this in mind, the first goal is to find a suitable cabinet or case for the rover station. This container must keep the electronics equipment dry but must also permit the flow of air used to ventilate the equipment; i.e., keep it from overheating. The first attempt for this box was a mailbox. Obviously, the mailboxes are designed to keep the mail dry, and they are certainly cheap enough and readily available at Home Depot or Lowe's Home supply. The entrances/exits for the cabling going to the rover rig where all placed such that they went into the bottom of the mailbox. A side benefit is that by bringing the cables into the bottom of the box forms a natural drip loop since the any cables from above the rig make a 180 deg. turn to enter the box. Cables coming from the lower side, such as a power cable from a battery, are shielded from above by the box itself. However, there is a problem with mail boxes which became apparent after a season or two for the rover rig configuration and when leaving the mailbox sitting outside at the base of a tower used for point to point communications. The mail box, while keeping out the rain out, did not keep out the insects such as spiders and Lady Bugs. An attempt was made to tape the opening seams with duct tape, but it was still impossible to keep all of the insects out.

A more viable solution was to use an ammo case shown in Fig. 1. These are very cheap, \$8 to \$20 depending on where they are purchased, and seal completely tight around the lid.



Figure 1 - AMMO BOX FOR RAMBO ROVER STATION

The top can also be completely removed to allow the installation and placement of subsystem components. On the bottom of the ammo box, a couple of small, 2 3/8th inch, muffin fans were mounted, redundancy in case of failure of one of the fans. A round, circular saw the size to cut the holes in the bottom of the ammo box was purchased from Lowe's Home Supply, and this was used to cut the holes for the fans as well as two more holes at the other end of the bottom of the ammo case to permit the entrance of air. These holes were covered with copper, window screen to keep out the Lady Bugs, spiders, and other insects.

These ammo boxes work very well, have a handle for easy transportation, are rugged, and inexpensive. For many folks in this area, they are simply left over from shooting sessions. See Fig. 2 below.

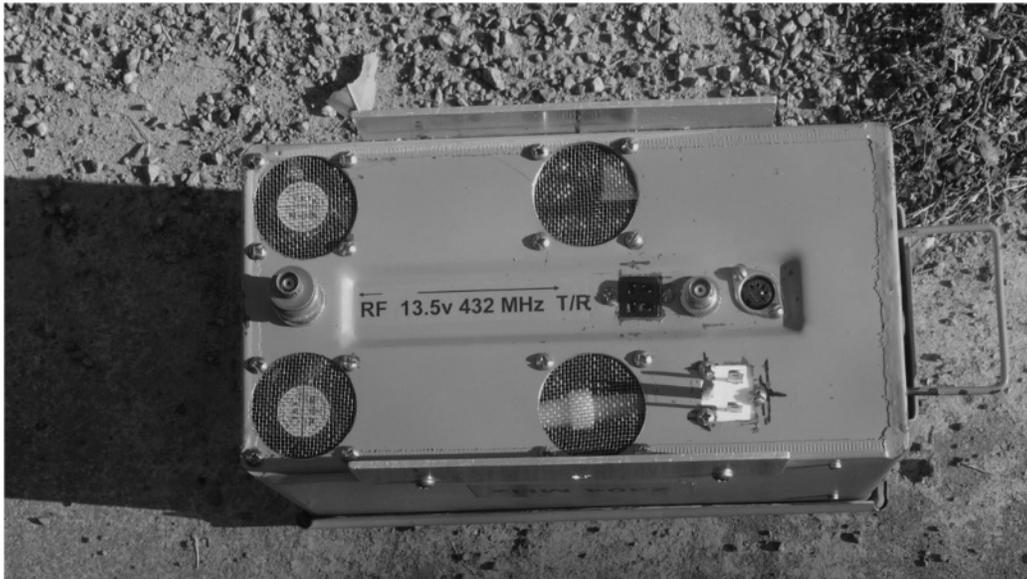


Figure 2 - BOTTOM VIEW OF ROVER CASE

The ammo box has two aluminum plates, available at the home supply store, attached to the sides of the box so that it can sit inside two angle aluminum rails on the side of a tower. This lets the rig sit at the base of the tower well above the ground level for home station use and also keeps the bottom of the ammo box above the ground when sitting the ammo box on the ground, keeping dirt and weeds out of the rig. The box has a Type-N connector for the antenna feedline, a four conductor Jones plug (or Power Pole) for D.C. input voltage, a BNC connector for the I.F., and a small connector for the key line for switching from transmit to receive. It also has a 120 V.A.C. plug in order that an A.C. extension cord can power the internal A.C. power supply for extended home station use or for those rare, almost non-existent occasions where A.C. power is available when roving. The four conductor Jones plug is wired so that it provide D.C. power to the rig when operating from A.C. power and disconnects the internal A.C. supply when D.C. power is applied to the Jones plug.

III. THE ROVER CONFIGURATION

The block diagram for the rover station is shown below in Fig. 3.

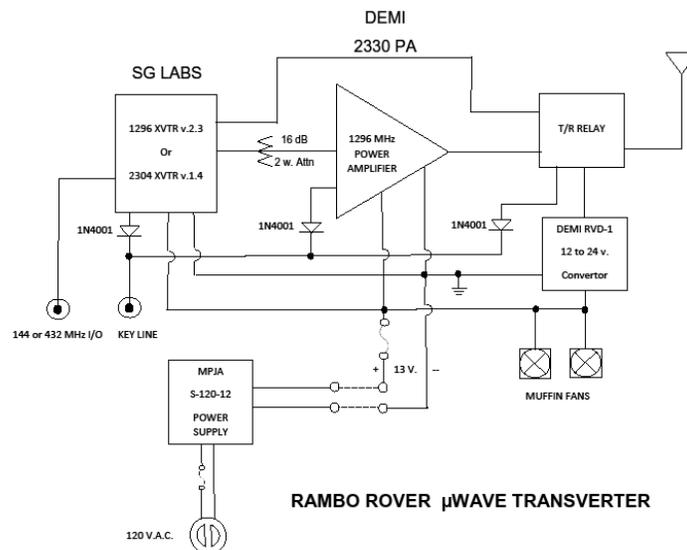


Figure 3 - RAMBO ROVER BLOCK DIAGRAM

For this system, the basic exciter used is the SG Labs 1296 V2.3 for 1296 MHz. A second SG Labs XVTR v1.4 was used for 2304 MHz, but it follows the same approach as that for the 1296 version built in another ammo box. The 1296 MHz version is described in this article, but both rigs used the same approach. The 2 watt output of the SG Lab exciter is 2 watts, and this is excessive as drive power for the DEMI 2330 power amplifier (PA) for 1296 MHz which was already on hand. To keep from overdriving the PA, an inexpensive 2 watt, 16 dB attenuator with SMA connectors was purchased from eBay. Since this rig is used on single sideband and CW, the 2 watt attenuator will accommodate the 2 watt output of the exciter without overheating as long as the key down time isn't excessive. One of the low loss, low cost readily available microwave relays is used for the T/R function, and a DEMI RVD-1 voltage converter is used to develop the 26 volts coil voltage necessary to pull in the relay while the 13 V.D.C. will keep the relay held keyed.

To isolate the various system components, classical 1N4001 diodes are used on the key lines for the SG Labs transceiver, the DEMI PA, and the change-over relay. Also of note is the MPJA (Marvin P. Jones & Asso.) switching power supply. This small, light-weight supply, S-120-12, is adjustable to 13.5 volts, supplies 10 amps of current, and does not appear to introduce additional noise to the 1296 receiver input. For a cost of ~ \$14, it negates the necessity to build a supply as the components cannot be purchased for that price. The four terminals on the Jones plug are shown to depict how either the A.C. supply or an external D.C. source, the battery, can be switched to power the rig from either house current or a battery.

IV. HOME STATION USE

For home station use, the decision was made to use the lowest loss coax available going up the tower but to keep the rig at the base of the tower. Tower mounted rigs are, no doubt, the most efficient configuration, but because of surge effects, the operator must be prepared to climb the tower fairly often or have a tower whereby the rig can be lowered to the ground. A good compromise is to use low loss coax up the tower, place the rig at the bottom of the tower where it can be accessed and maintained, and run the I.F. line inside the shack to cut down on RF losses. This may introduce an additional 1 or 2 dB feedline loss in front of the receive preamplifier, but keep in mind that the antenna is looking at a nominal 290 deg. K propagation median and not the 50 deg. K as typical with EME. As long as the receiver system noise temperature is less than 290 deg. K, the placement of the rig at ground level is o.k.

A tower tray is constructed from angle aluminum by attaching two pieces of angle aluminum to the tower legs extending out from the tower as support arms. On top of these support arms, two more pieces of angle aluminum are spaced such that the standoff rails attached to the ammo case fit within the angle aluminum shown in Fig. 4. Fig. 5 shows the standoff rails attached to the side of the ammo case. These rails also keep the bottom of the ammo case off of the ground and out of the dirt when moving the rig from the car to the tower or other operating position.

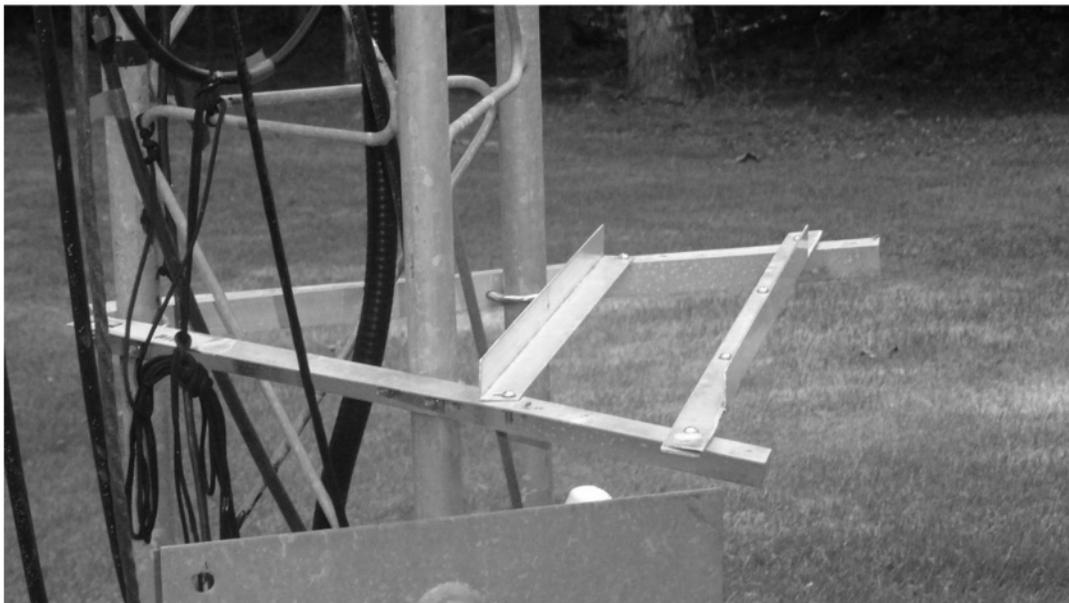


Figure 4 - TOWER MOUNTING ARMS

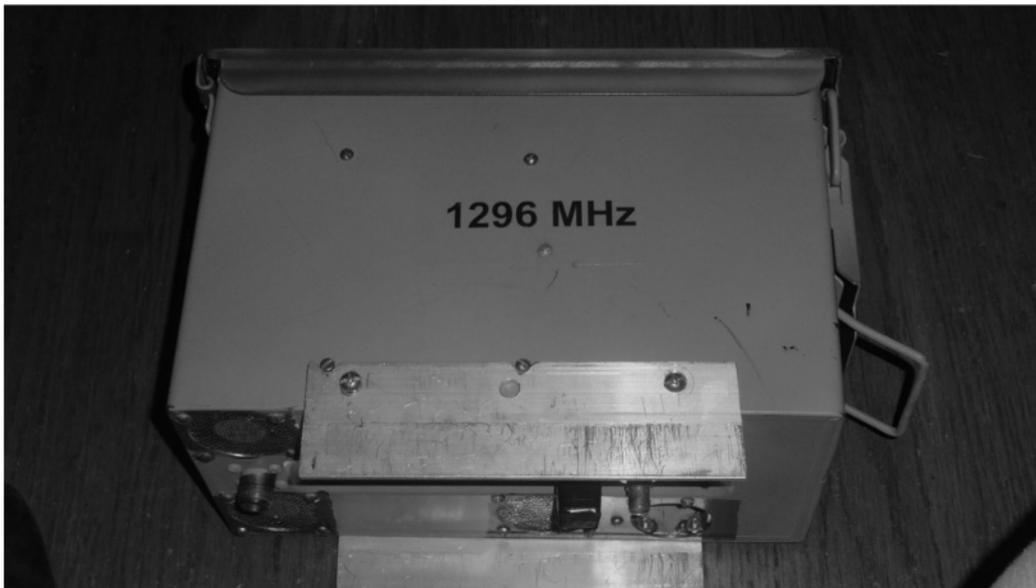


Figure 5 – SIDE RAILS ON CASE

The ammo case sitting in the tower mount is shown in Fig. 6. This arrangement provides a natural “drip loop” for cabling in and out of the transverter to keep water out of the cables and connectors.



Figure 6 - RIG ATTACHED TO TOWER

V. AUXILARY EQUIPMENT

Two antenna configurations are shown, one for the roving and one for the home fixed station. A small antenna that a compliments a multi-band, microwave rover station is the Vivaldi antenna. While there are several manufacturers of Vivaldi antennas, a version by RF Space which is useful in that it provides adequate gain, handles up to 50 watts of power, an acceptable SWR, and covers all of the popular microwave bands. This antenna is shown below in Fig. 7.

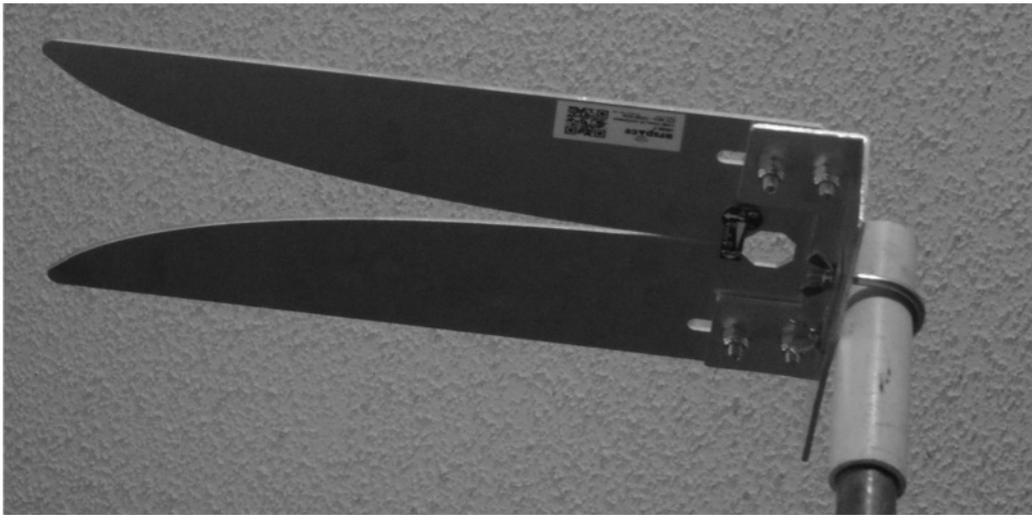


Figure 7 - VIVALDI ANTENNA, 1.2 TO 12 GH

For home station use, a dual band BBQ grill, dish reflector will provide excellent performance for both 1296 and 2304 MHz shown in Fig. 8. This antenna uses a Log Periodic Feed by placing the phase centers of both bands at a position to straddle the focal point of the BBQ grill reflector. The BBQ grill reflectors are available as WiFi antennas or short range data link antennas usually for 2400 MHz. The measured gains for this antenna are 16 dBd for 1296 MHz and 21 dBd for 2304 MHz. Both gains are approximated 1 dB below that achievable for a 2 ft. x 3 ft. reflector optimized for single band operation with the feed phase center at the focal point, but that's the price of compromise for one antenna to cover both of these bands. The one antenna is certainly easier to transport and setup than two of these antennas, one for each band.

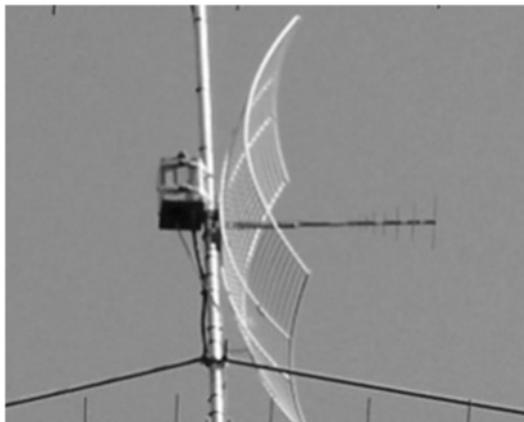


Figure 8 - 1296/2304 MHz ANTENNA FOR HOME USE

Power for the rover rig is provided by a Suaoki Li-Ion battery. This particular 13 volt battery provides 18 amp-hours of energy, 300 amps of peak current, and weighs ~ 3 lbs. There is also an additional low current output which can be set to 5, 9 or 12 volts to provide power for a 10 MHz TCXO or other reference source. It also has a couple of USB outputs for USB powered devices. Batteries similar to this one are now available through the Wal-Mart website to be delivered to a nearby store for pickup.

The newer version of these batteries also include an internal fuse to prevent catastrophic results (explosion) should the terminals accidentally be shorted. The battery is shown in Fig. 9 on the following page. To say the least, this battery is considerably easier to transport to the top of a tower than a deep cycle, lead acid battery!



Figure 9 – SMALL, LIGHT WEIGHT BATTERY FOR ROVING

VI. CONCLUSION

Roving, especially microwave roving, is extremely interesting, challenging, and rewarding. The recent introduction of modern subcomponents permits the microwave ham to configure an efficient station that isn't too bulky to take along, and it provides the opportunity to have some exiting QSO's such as my most rewarding operation a couple of year ago during the 10 GHz contest weekend when WA5YWC was operating on Grand Isle, south of New Orleans, and I was set up on Mexico Beach, some 25 miles S.E. of Panama City, FL. That particular QSO across the Gulf of Mexico was at a distance of 285 miles, my best DX on 10 GHz to date. The only factor that prevented QSO's of 600 to 800 miles was the lack of activity down the South Texas coastline.

A Method to build your own Waveguide to Coax Transition

Bob Stricklin N5BRG
N5BRG@ARRL.net

To operate in the microwave bands you typically use an antenna system that requires converting from coax to waveguide. In this paper I will refer to these devices as converters but they are also known as transitions and adapters. Converters can be found in flea markets and on eBay. If you find converters they typically range in price from \$20 to \$250. A few examples of converters available on eBay, as of this writing, are shown in Figure 1.

	<p>WR-90 WR90 - SMA WAVEGUIDE ADAPTER Pre-Owned</p> <p>\$93.75 or Best Offer +\$1.50 shipping</p> <p>From Greece</p>
	<p>10-11Ghz WR-90 to SMA Waveguide to Coaxial Adapter 68B62-003 New - Open box</p> <p>\$209.00 Was: \$237.50 or Best Offer +\$2.50 shipping 12% off</p> <p>From Greece</p>
	<p>WR-90 WR90 TO SMA WAVEGUIDE TO COAXIAL ADAPTER QUASAR Pre-Owned</p> <p>\$82.50 Was: \$93.75 or Best Offer +\$1.50 shipping 12% off</p> <p>From Greece</p>

Figure 1

When you make a purchase of one of these you are sometimes surprised to find the part is advertised as WR90 but the dimensions of the waveguide or the mounting holes do not match other WR90 waveguide pieces you have. In addition you have no idea what the real performance of the converter is until you test the device.

The purpose of this paper is to detail a method of building your own converters using readily available materials. If you have access to surplus waveguide this can be used as well. In some cases the inside surface of the waveguide is plated usually with silver and this will result in an improvement in performance by a few tenths of a dB. If you want to consider plating your own waveguides I have added a couple of references about plating. If you're near an industrial area you can find a business that does silver plating.

Paul Wade, W1GHZ published an excellent article (1) on building converters which covered his experience and gives details on an approach which may be used to build a microwave converter. I used Paul's article to help guide me in my efforts. Paul's approach was to start with a piece of waveguide made for the target band, WR 90 for 10 GHz as an example. If you want to follow Paul's approach you might look for waveguide or waveguide parts like the one shown in Figure 2 found on eBay. Then you could cut the waveguide section into two pieces and place a plate over the end of each of the pieces and add SMA connectors. The result would be two converters. Since you inherit the flanges from the existing piece this minimizes machining and effort. Finding an economical part to start with may be difficult.

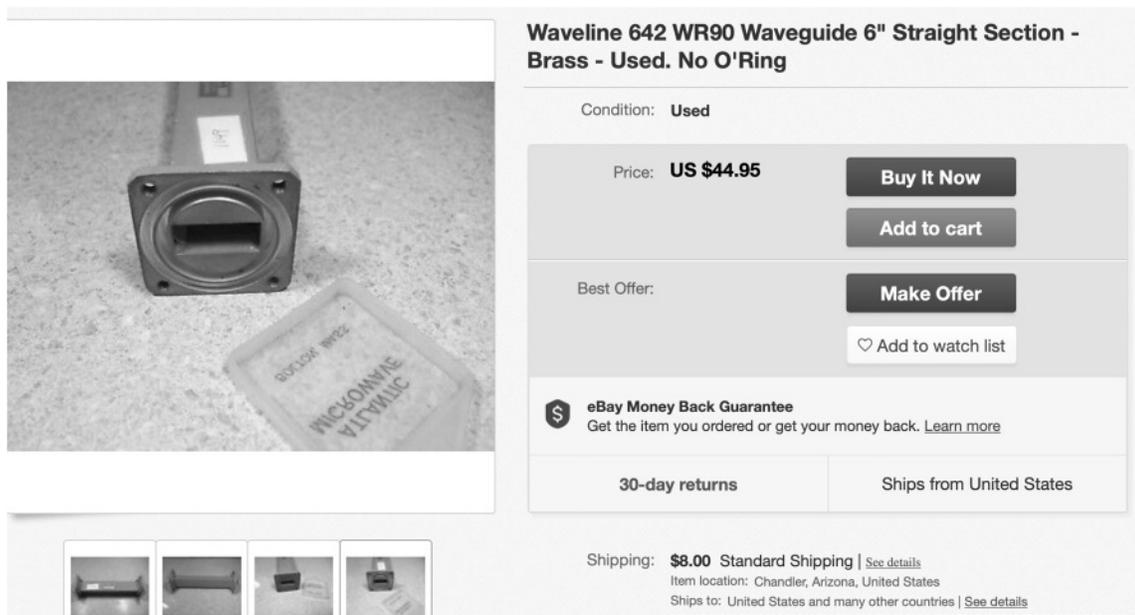


Figure 2

Another approach to this is to use brass rectangular or square tubing and make your own flanges. Brass is chosen because it can be soldered. There are metal distributors like McMurry Metals in Dallas Tx (2) or OnlineMetals.com (3). Selling various sizes of brass tubing, bar stock and other shapes. You can likely find a distributor in a city near you if not you can order the material on line.

To determine what size tubing will work for a particular band I studied David Pozars' book Microwave Engineering (5). In Section 3.3 he discusses Rectangular waveguide giving the detail formulas and physics involved. He also covers circular waveguide. From this reference I developed an Excel spreadsheet to calculate the cutoff frequencies of various rectangular tubing sizes. By using the Excel spreadsheet as a guide I was able to consider the standard sizes of brass tubing made in industry today. Table 1 shows my results with some of these sizes along with the cutoff frequencies and the examples from the Pozar book. The goal is to select tubing with a cutoff frequency below your desired operating point. The upper limit of the tubing will be two times the cutoff frequency. At two times the cutoff frequency multiple emission modes can occur.

	A	B	C	D	E	F	G	H	I	J	K	L	M	N	O	P
1	Wave Guide Cut off Frequency						c=	Wavelength								
2							299,792,458.00	Free Space								
3								Waveguide								
4	Rectangular Waveguide	TE10														
5		(in)	(in)	(m)	(m)	(GHz)							10.368GHz	0.02892	0.03836	1.14
6		a	b	a	b	Fc							1/2 Lamda	0.01446	0.01918	0.57
7		0.421259	0.4	0.011	0.010	9.71		Example from Pozar Pg 117					1/4 Lamda	0.00723	0.00959	0.28
8		0.9	0.4	0.023	0.010	6.56		8.74								
9		1.122	0.497	0.028	0.013	5.26		7.01								
10		0.622	0.311	0.016	0.008	9.49		12.65								
11		0.875	0.375	0.022	0.010	6.74		8.99								
12																
13	Round Pipe	TE11	Radius													
14		a	a	P'11=	1.841											
15		(in)	(cm)													
16			0.5			12.19		Example from Pozar Pg 127								
17		0.375	0.95			9.23		12.30								
18		0.5	1.27			6.92		9.23		12.30						
19		0.75	1.91			4.61		6.15		9.23						
20		1	2.54			3.46		4.61		6.15						
21		1.25	3.18			2.77		3.69		4.61						
22		1.5	3.81			2.31		3.08		3.69						
23		2	5.08			1.73		2.31		3.08						
24		3	7.62			1.15		1.54		2.31						
25																

Table 1.

I have included the Excel file in the MUD 2019 data disk. If you would like to determine frequency cutoff for additional sizes of rectangular or circle tubing it is easy to expand the Excel file. You could also look at cutoffs of standard sizes of WR waveguide like WR90 which is 1.00" X 0.50" with wall thickness of 0.050" making inside dimensions 0.90" X 0.40".

After you have selected a tubing size that will propagate your wavelength you will need to build the proper size flange. My initial goal was 10 GHz and I will explain the process used to build a converter for this band. I purchased flat pieces of brass stock cut from a strip of 1 5/8" X 1/8" bar stock. The pieces received were 1.68" X 1.68 X .132". I also purchased a 2 foot length of rectangular brass tubing with outside dimensions of 1.0" X 0.5" and wall thickness was specified as 1/16". So the inside dimensions of the waveguide will be 0.875" X 0.375". The dimensions of the tubing I received was actually 0.875 X 0.380 X 0.058".

The distributor sheared the flat stock into 1.68 by 1.68-square pieces for me. The next step to make a flange is to cut the slot in the center and add the holes. I also used a file to round off the sheared edges and to break all the edges and corners around the outside of the part to make the flange. The best way to get the slot is to have them machined in by someone that has a milling machine. If this is too costly or difficult you could mark the slot and drill holes inside the area that must be removed and using a small flat file remove the material so the rectangular tubing will fit into the slot. I have included a drawing, Figure 13, with dimensions for a finished flange that will fit 1" X 0.5" brass tubing. If the slots are

machined there is typically a radius (end mill radius) in the corners. This will need to be filed square or the corners of the rectangular tubing will need to be rounded to fit into the flange.

For the rectangular tubing I used a miter guide and a small tooth saw to cut the tubing to length. Then I marked drilled and tapped the holes needed in the tubing. The finished part needed is shown in Figure 11. The length on my tubing/waveguide was to help me handle it and mount later in front of a dish. The length can be short as long as you or more than one wavelength in front of the probe on the flange end. At high power there can be some internal surface resistance causing a small loss but we are typically not using so much power.

From brass bar stock I purchased which was 1" X 0.375" I machined a plug that would slip into the rectangular tubing. The finished drawing for this is shown in Figure 12. A hole was drilled and tapped at 2-56 in the back of the plug to add a screw to use as a handle for adjusting its position and tuning.

If you study Paul Wades paper and Table 4 in Paul's paper you will see the back wall of the WR90 converter ends up so close to the insertion probe that you may hit any screws that protrude inside the cavity or burrs from the holes drilled and tapped for the screws. To avoid this, clean away any burrs and be careful with your screw length. Try not to scratch up the inside of the tubing more than necessary.

The next step is to solder the front flange on the rectangular tubing. You can do this by lightly burnishing the surfaces that will receive solder with some fine sand paper or scotch bright. Then insert the tubing in the flange and using a flat surface make sure the front face of the tubing and the flange is well aligned. Using a square make sure you can get the flange face to be perpendicular to the length of the tubing. Then wrap about 8 turns of .025" diameter solder (adjust the turns for the solder size you have on hand) around the tubing and keep the solder close to the backside of the flange. The process I used was to double check the flange surfaces perpendicularity then place the assembly on a hot plate I purchased at Walmart for less than \$20. Next I placed a bead of flux (Chipquik SMD291) around the solder. Then turn on the hot plate at a high setting and let it begin to heat up. The flux will smoke so do this in a ventilated area or outside. When it gets hot enough you will see the solder wet and flow. Some of the solder will wick down in the space between the flange and tubing and make sure you have enough solder to fill the void in the mating space. Add more solder by hand if needed. This can be a bigger issue if you hand milled the slot.

Don't touch anything while the solder is molten and as soon as you see the solder has melted well you can turn off the hot plate and let things cool down. If you must touch the assembled waveguide wait until the solder solidifies then use plyers or something to pick it up. Don't burn yourself and wear some safety glasses!! Several photos are shown in Figure 3 illustrating this process.

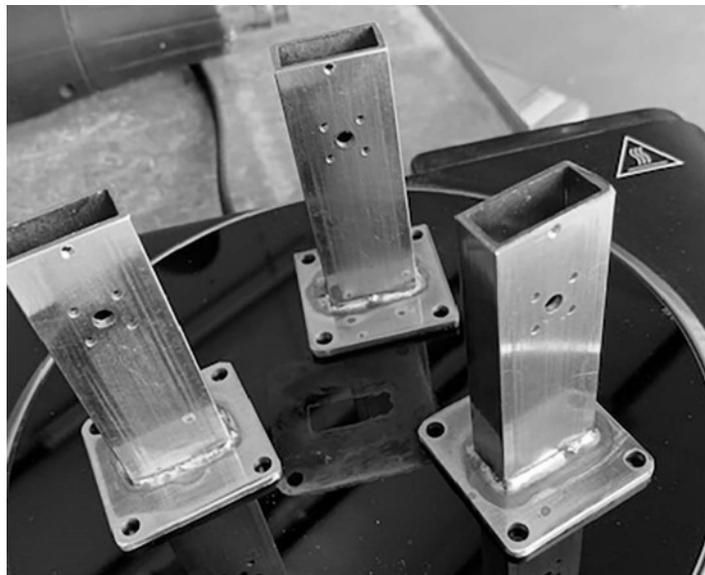
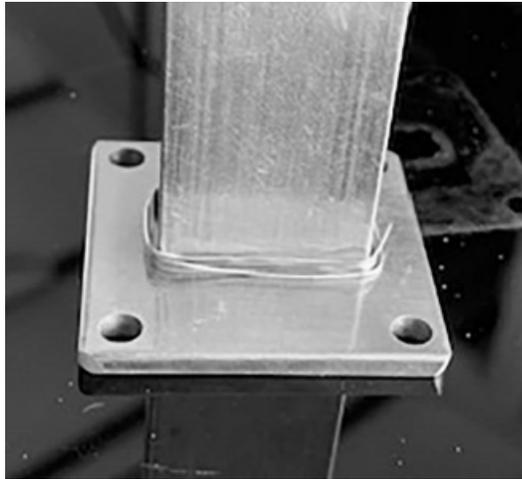


Figure 3.

When things have cooled down you should clean the part using a tooth brush or similar and some water or alcohol. I used some Q-Tips to clean the inside of the tubing. An alternative for cleaning would be to dip the part in toilet bowl cleaner for a few seconds and then clean the part thoroughly with fresh water. The bowl cleaner is a mildly acidic and it will shine up the brass.

The next step is to prepare the insertion probe on the SMA connector. If you can find an SMA connector with a 0.625" diameter pin more than 0.30" in length this will be ideal. If not, you may need to purchase some .0625" mm round brass tubing and solder this on a SMA connector that has a pin diameter which will fit inside your purchased round tubing. The goal is to get the pin to a length of 0.285" with a small flange of Teflon that is 0.056" extending from the base of the SMA. I have a drawing showing the finished SMA with probe in Figure 4.

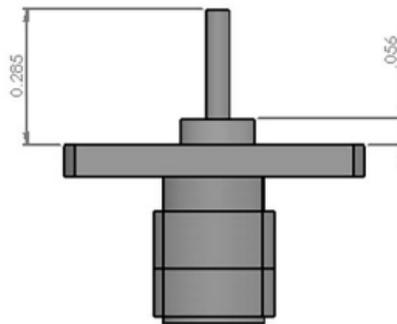


Figure 4.

Apply the probe to the waveguide and secure it with screws. Make sure the plug will slide into the backside of the waveguide freely. The tuning plug which will become the back wall of the waveguide. This plug must go down into the waveguide so the front face of the plug is very near the probe pin. You will position this by holding the screw placed in the back of the plug. The conductivity of this plug to the walls of the waveguide is critical. For adjustments you are using low power levels and may see some variation in readings but you can determine the best operating point. Later the plug will be soldered in place and conductivity will be good and support high power.

For testing and adjustment I attached a waveguide dummy load, purchased on eBay, to the flange. You could use something like this or you might attach another waveguide converter with a 50 ohm dummy load attached to the SMA connector. For a signal generator I used an HP8350A with plugin 83592A set to scan from about 9.5 GHz to 11 GHz and output +7 dBm of power. Then I monitored the reflected power using a directional coupler and a HP853A with plugin 8559A spectrum analyzer. I show this arrangement is Figure 5 block diagram and the photo in Figure 6. I placed a 3 dB attenuator in front of the directional coupler to be sure the signal generator would see a 50 ohm impedance load. I also used a 6 dB attenuator in front of the spectrum analyzer to protect the input and to make sure again that I had a 50 ohm impedance.

The test equipment shown was used because I have it available but you should be able to tune up the converter with a 10 GHz power detector of some sort to detect the power coming from the directional coupler and a W1GHZ personal Beacon setup for 10.304 GHz. You may also just leave the flange end open to the air and tune for minimum reflected power.

You will see my parts in the photos are machined for a SMA connector with four screw mount holes. I have included this drawing in the drawing files if you want to use this type of SMA. The two-screw mount SMA will be better for staying away from the tuning plug. This may depend on the SMA connectors you can find.

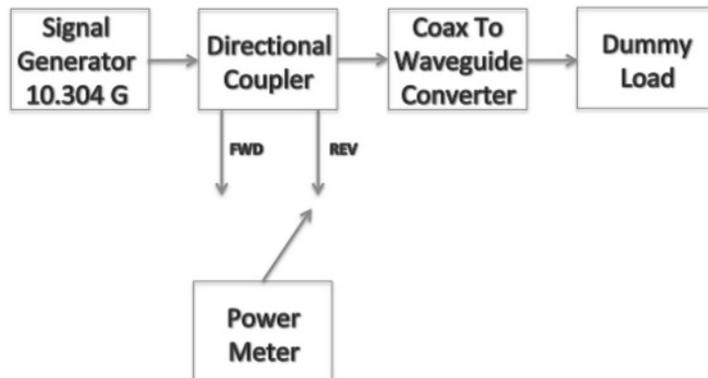
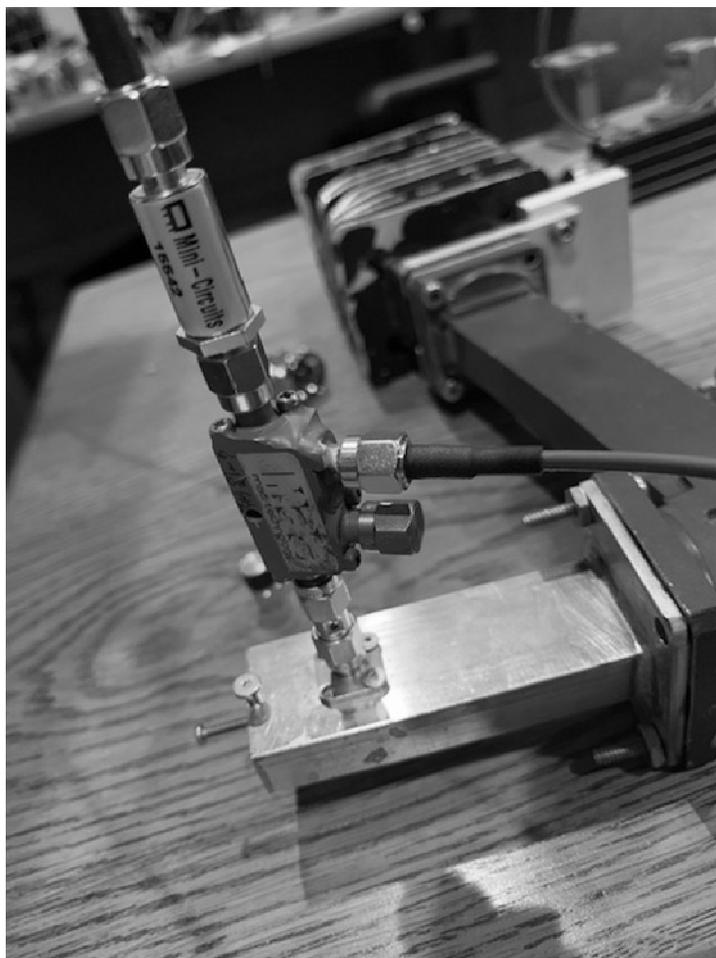


Figure 5



Setup with converter dummy load and directional coupler.

Figure 6

With everything connected I proceeded to adjust the location of the plug in the back of the waveguide looking for the lowest possible power level for the reflected power. Note my photo in Figure 6 has the directional coupler configured for forward power measurements. It's important to measure the reflected power first so the directional coupler would be reversed. When I found the lowest reflected power level I tightened the holding screw on top of the waveguide down against the adjustment plug to lock the plug in place. Then I rechecked the power level to be sure the power remained the same. Very small movements can make a lot of difference in the process so you have to just be patient and carefully adjust things.

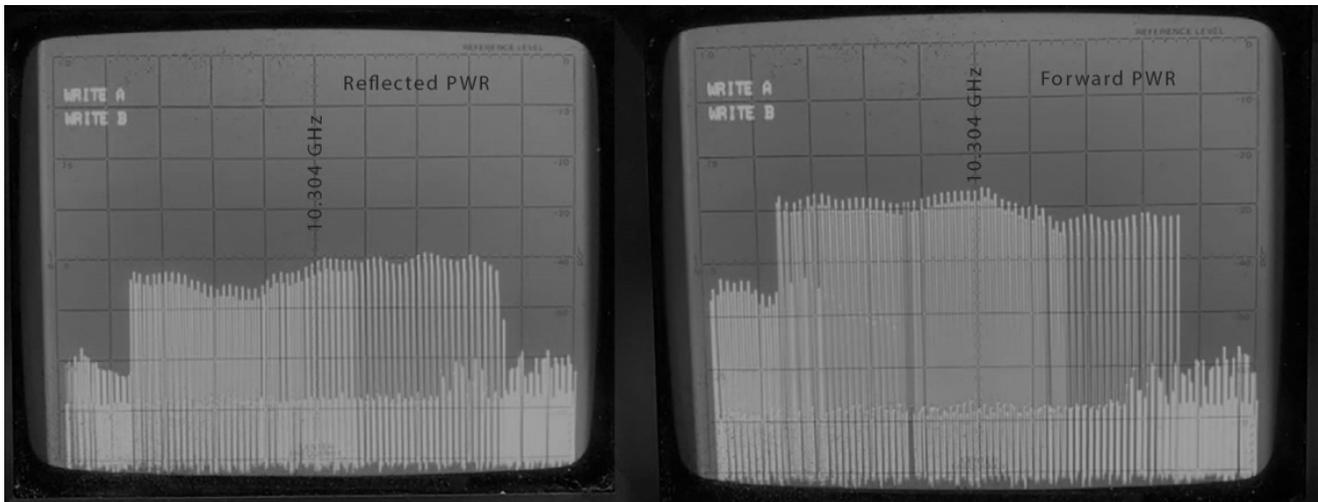


Image of HP853A/8559A with center at 10.304 GHz and 200 MHz per cm.
Figure 7

Next I reversed the directional coupler and measured the forward power delivered into the converter. With the reverse power measured in dB and the forward power measured in dB I can calculate the SWR. My results were a power difference of 12 dB which according to a table I found (3) is an VSWR of 1.68.

I then measured the distance from the back of the plug to the back end of the waveguide. With this measurement and knowing the dimensions of my plug I could determine the distance from the probe pin to the effective back wall of the waveguide. I was impressed to find the distance was 0.125 inches, which was very close to Paul Wade's table 4 numbers shown in his paper.

To check my setup I inserted an HP X281A converter I had and measured its VSWR and found it to be 2.0 similar to the one I just built. This was 2.0 using the waveguide dummy load. With the dummy load an ideal match would read very low SWR.

I should be able to improve the SWR of the finished converter by adding a tuning screw between the probe and the waveguide flange. This screw would add capacitance to the converter. If I tune with the screw using a waveguide dummy load, which should have an impedance of 377 ohms, I can lower the SWR more. I may be able to reduce the SWR to near 1/1, which would be an improvement in the signal of 0.47 dB.

Now that I had this working I tried different connectors and probe lengths but did not see any significant improvement in performance. In most cases it was worse. However, this process showed me that there

might be a way to improve performance later by better targeting the converter physical dimensions for a frequency of 10.304 GHz. This may be a subject for a later paper.

The last step was to apply solder to the plug to lock the plug into position and seal the back of the converter. I applied a bead of the same flux used earlier around the back parameter of the plug. Next I used a 1500 Watt Dual Temperature Heat Gun I purchased from Harbor Freight for about \$15 to heat the part hot enough to flow solder. I place the part on a metal table and propped it up using a large diameter screw head while heating. Important not to hold the part with a vice or something that will sink heat away. When the plug and the waveguide were hot enough I hand feed solder around the seam and filled it in. Turned off the heat gun and let the part air cool.

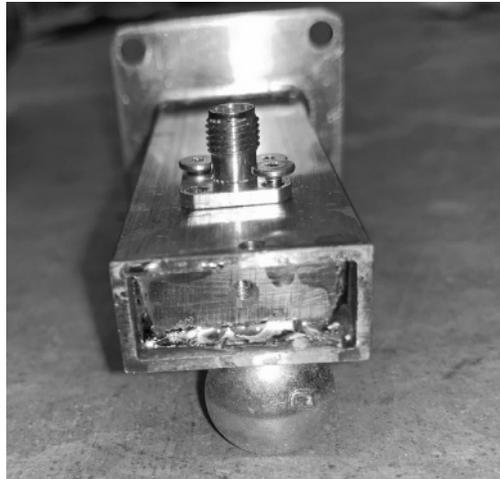


Figure 8.

Finally, I retested the part and verified its performance. The waveforms from testing the final part are shown in Figure 7.

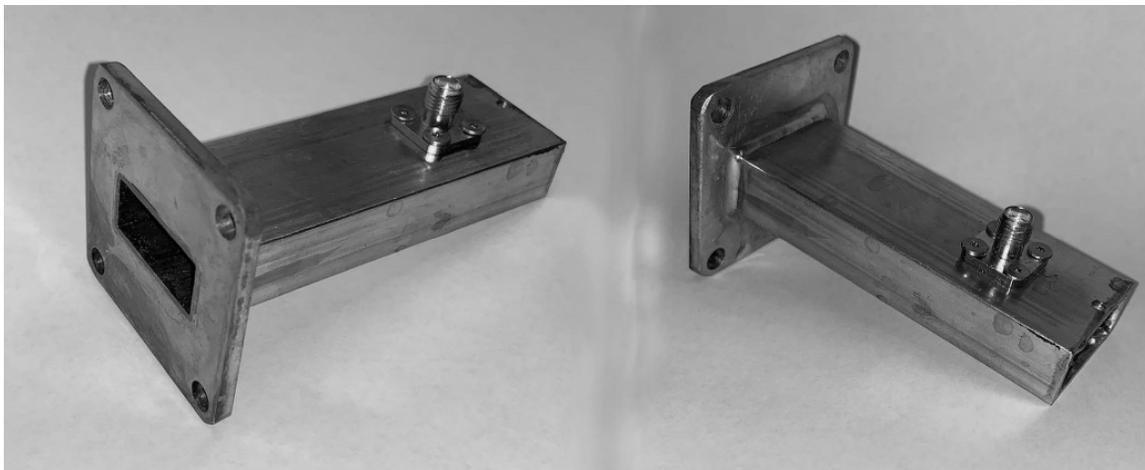


Figure 9.

I enjoy building things like this and I think others do as well. I have found several converts are needed for dish antennas and testing as I expand my operating capability and knowledge in the microwave

bands. I will try to build converts for several more bands and hope to improve on their performance as time permits.

All the drawings needed to build the 10 GHz waveguide shown here are included in the distribution thumb drive for MUD 2019. Multiple cad file formats and PDF files are included.

Special thanks to Harold Reasoner, K5SXX for reviewing this paper and the discussion and suggestions that resulted.

References:

1. http://www.w1ghz.org/QEX/Rectangular_Waveguide_to_Coax_Transition_Design.pdf
2. <https://www.minicircuits.com/app/DG03-111.pdf>
3. <https://mcmurraymetals.com>
4. <https://www.onlinemetals.com>
5. "Microwave Engineering" David M. Pozar 3rd or 4th edition ISBN-10: 0470631554
- 6 https://www.qsl.net/w1ghz/small_proj/small_proj.htm

Plating Waveguide: https://library.nrao.edu/public/memos/temp/Petencin_Plating_Inside_05.pdf
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<https://www.finishing.com/faqs/silverathome.shtml>

Appendix:

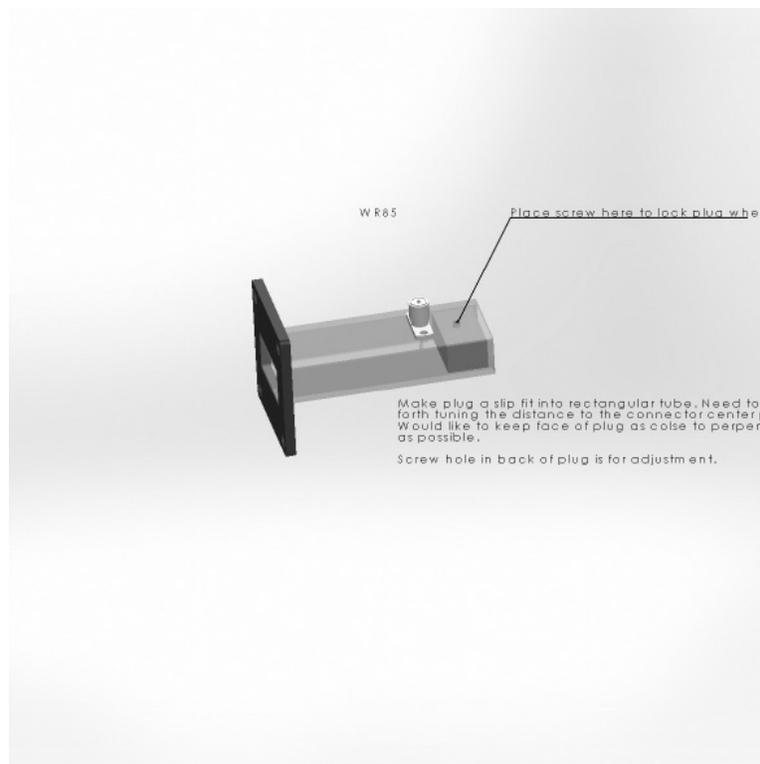


Figure 10.

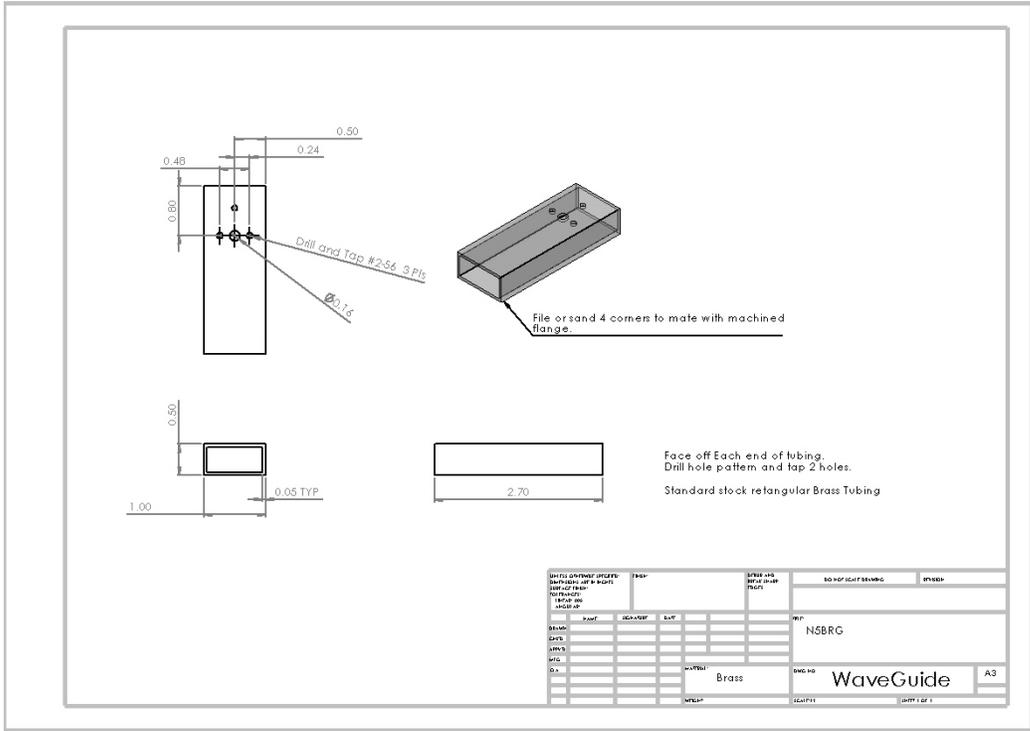


Figure 11.

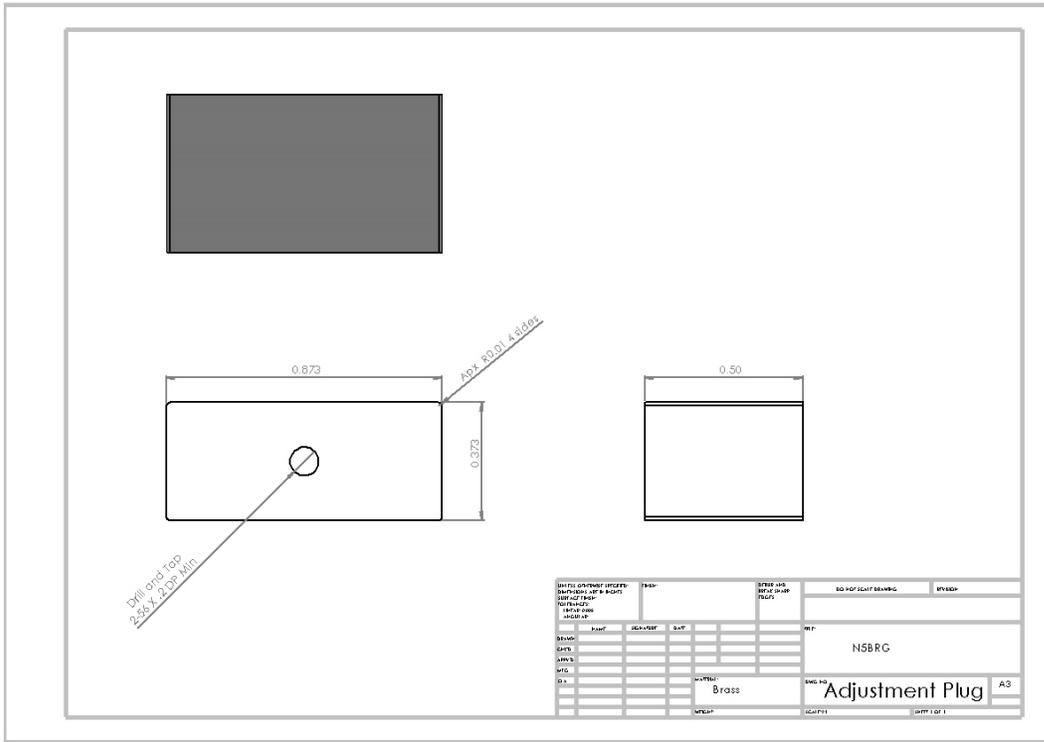


Figure 12.

The Six Centimeter Feed Experiment

Skip MacAulay, VE6BGT

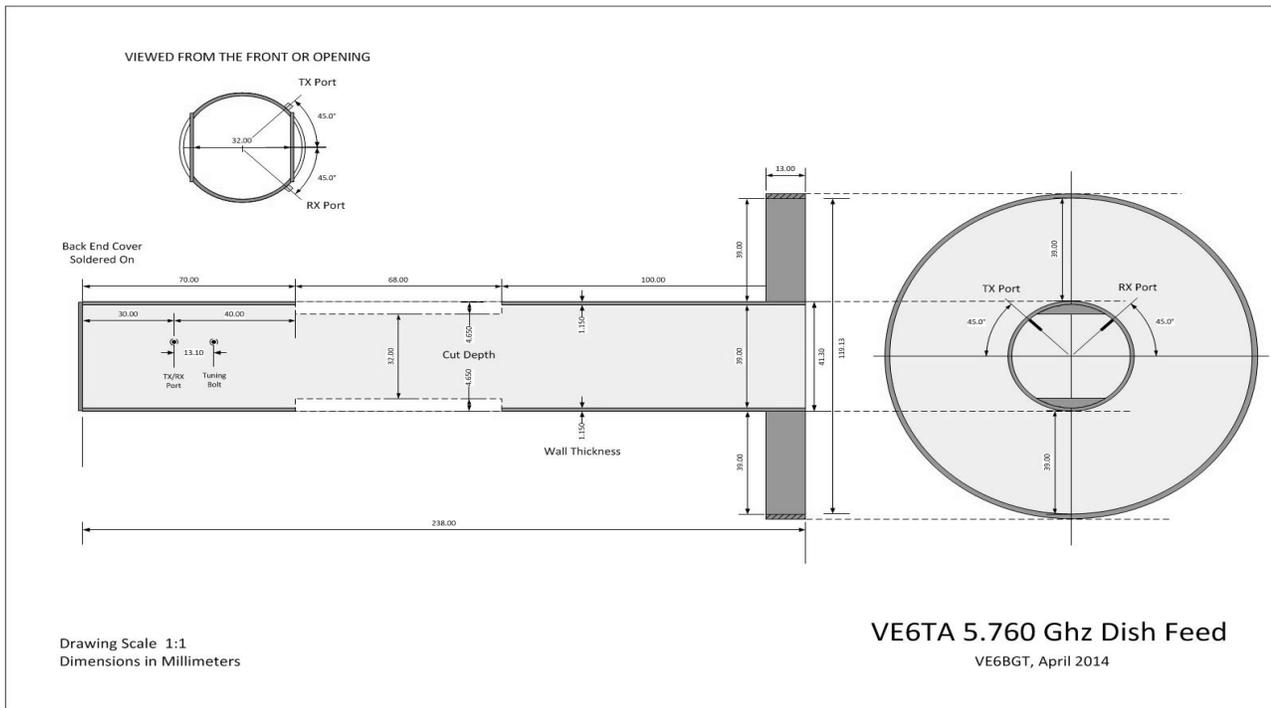
After building successfully the 100 watt Gan FET amplifier, the next step in the 5.760 Ghz experiment was to build up the dish feed assembly. This was going to contain the complete transvertor, amplifier, preamp, and feed ant all on one mounting plate to be placed up on the dish feed mounting shelf. The first part was to build some kind of feed ant. I had built a “squeezed” tube feed a few years earlier for Grant, VE6TA which turned out well. So with that design as one and then scaling down the more common VE4MA bolt polarizer style, I came up with three types or models to experiment with.



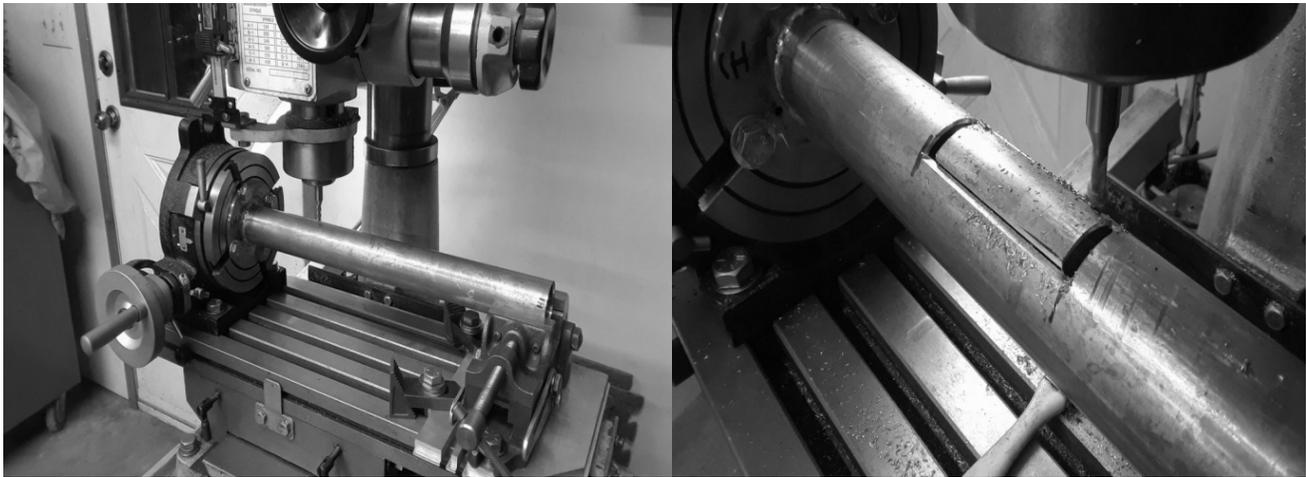
The Three Amigos

I decided to build the squeezed tube one first and tune, then the scaled down bolt polarizer one which made the adjustment bolt placement quite close together. I thought just for experimentation what would happen if I made the copper pipe longer and spaced the polarizer bolts farther apart? So a second one was drawn up and built.

The launcher sections I have used up till now on lower frequencies have always been the scalar ring style. I decided since at this higher frequency and the f/d ratio of my dish being .444, I wanted to try out a W2IMU style of launcher. The same IMU was used at first for all three feeds.



The Original “Squeezed” Feed Specification Drawing

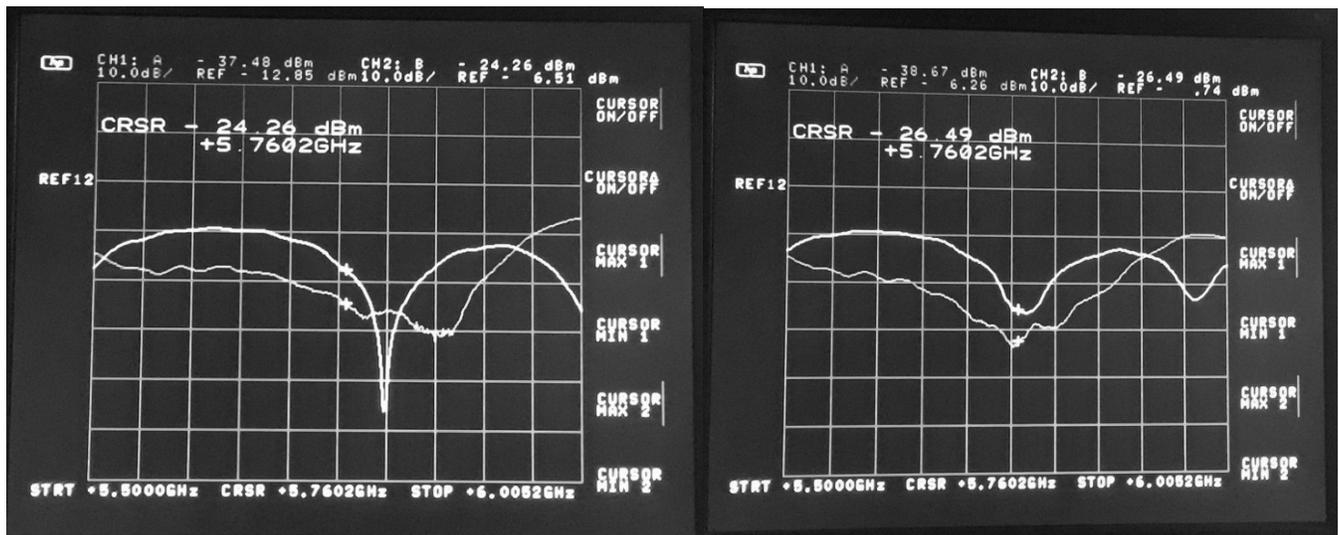


Cutting the CP Holes

The squeezed feed was built from copper pipe with an outside diameter of 41.3 millimeters, inside at 38.5 millimeters. With the copper pipe attached to a rotary table I could take great detail in cutting the two slots or rectangular holes. Then when the copper plates were soldered into the holes the inside spacing between the two plates would have a distance of exactly 32 millimeters. This space is for proper creation of the signal for circular polarization.



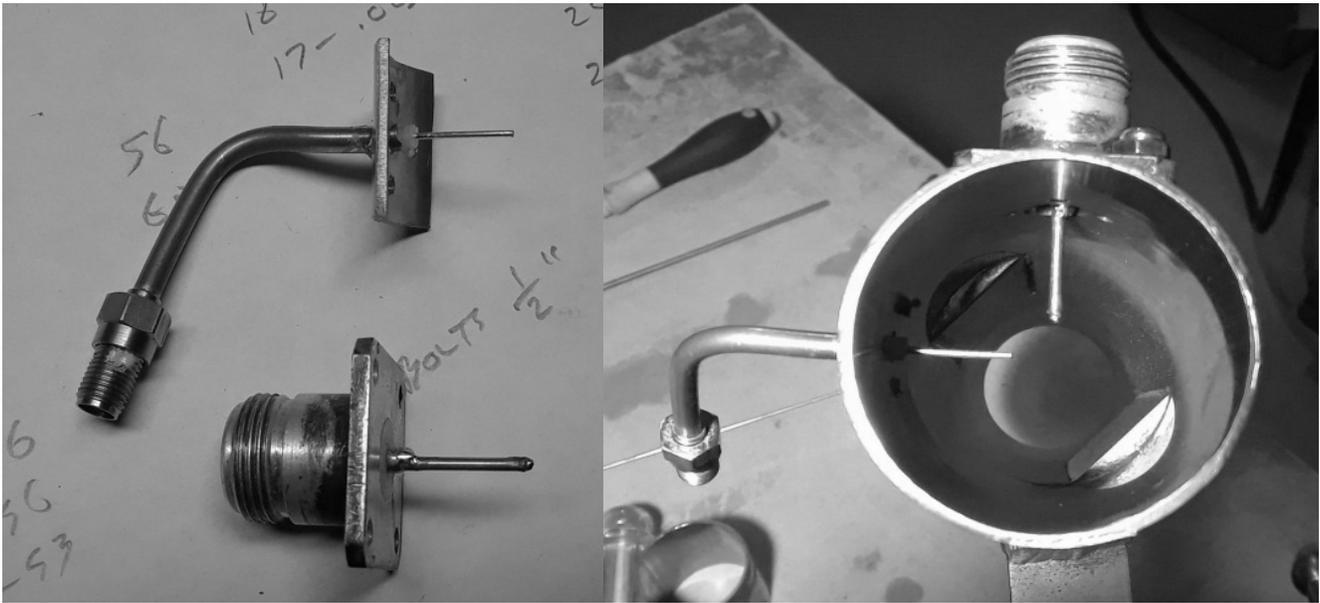
This picture shows the finished squeezed feed with the two plates soldered in and the new IMU launcher section. There is also an adjustable sliding back wall or plate to aid in tuning. I planned on using a SMA connector for the transmit probe only and usually a custom made receive connection made from a piece of .141 semi rigid coax to connect directly to the isolation relay, eliminating one connector for less line loss. I decided I would use the SMA connectors for both at first for the testing.



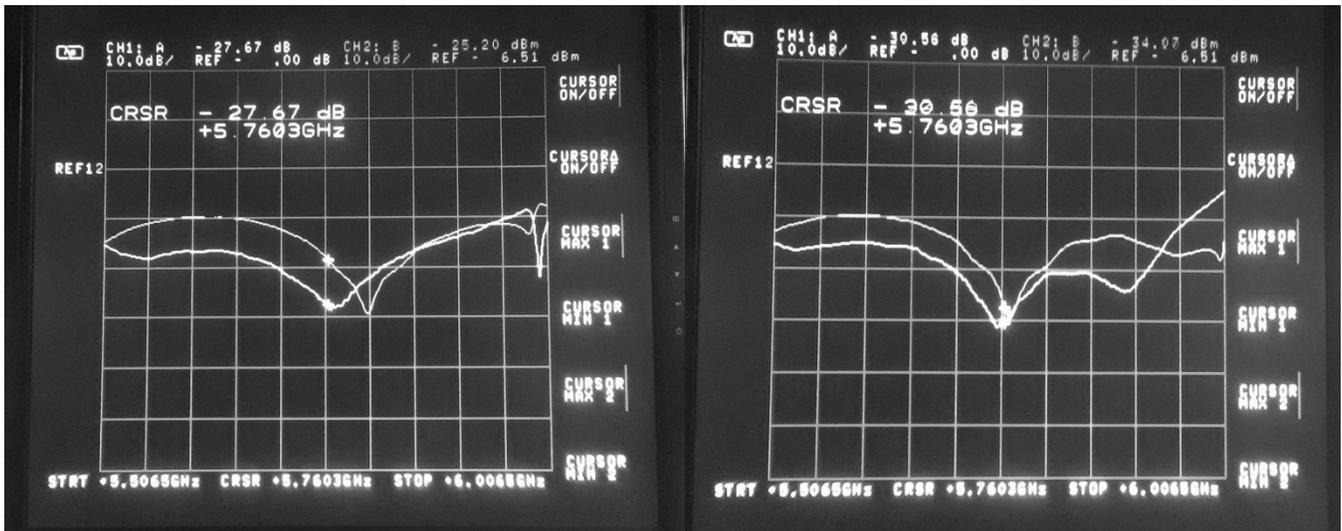
TX Probe Tuning.

Back Plate and IMU Launcher Adjustment

These two pixs are the first results of tuning the probes of the squeezed feed tube showing the return loss of TX probe and isolation to the receive probe. The left pix is of just the probe lengths adjusted, the RL is at 25 dB, isolation at 18dB. The right pix is after adjusting the back plate and IMU launcher. Doing this made the resonant dips move right on frequency and help the RL to 32 dB, ISO at 25dB. But the Circular Polarization test came up weak, as much as 5 to 6 dBm difference in amplitude. There is more on this CP test later.



I then wanted to convert the feed over to an N connector for the TX probe and a custom rigid line connector for the RX probe. This RX probe line hooks directly to the SMA Isolation relay.



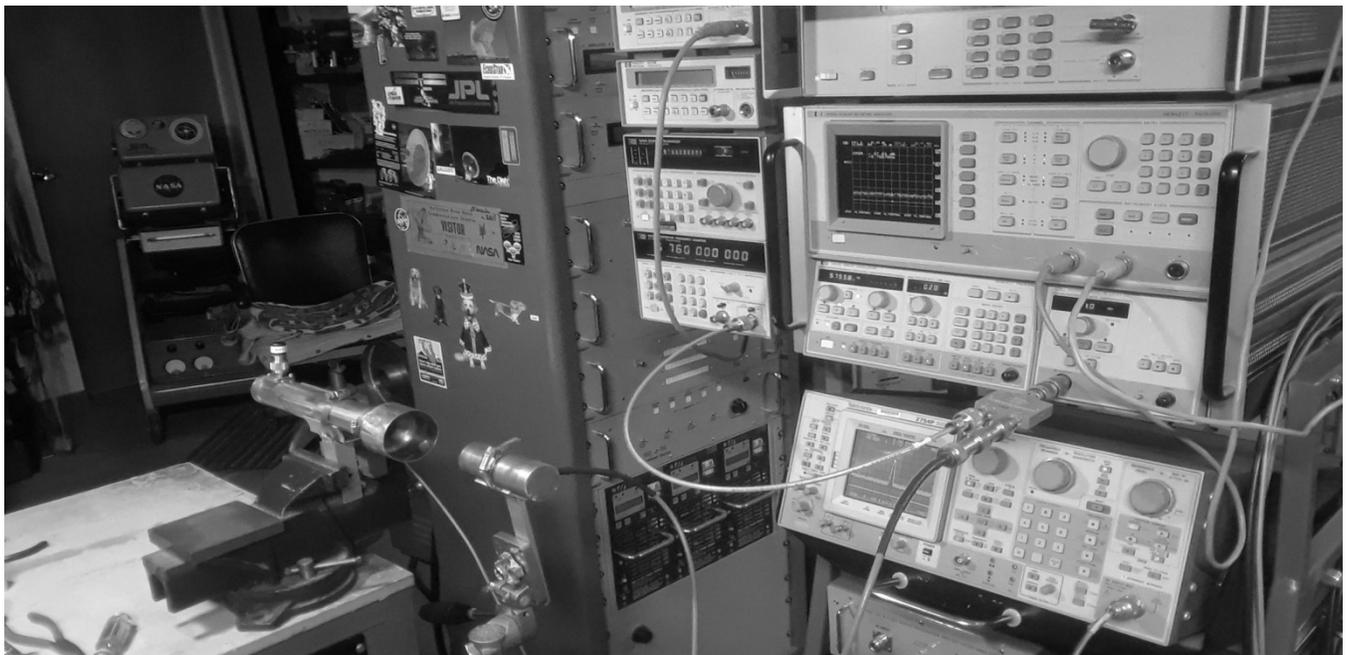
First Sweep: RL = 27 dB, ISO = 19 dB Best Adjustments: RL = 30 dB, ISO = 28 dB

After the new connectors were added the probes were swept and tuned again. The pix on the left are the probes adjusted and the back plate tight up against the body, the IMU launcher is completely off. The right pix is the best of all adjustments, the back plate moved out 1 mm, the IMU installed and slid out 4.5 mm. One thing I really noticed was the accuracy of construction the IMU launcher needed to be.



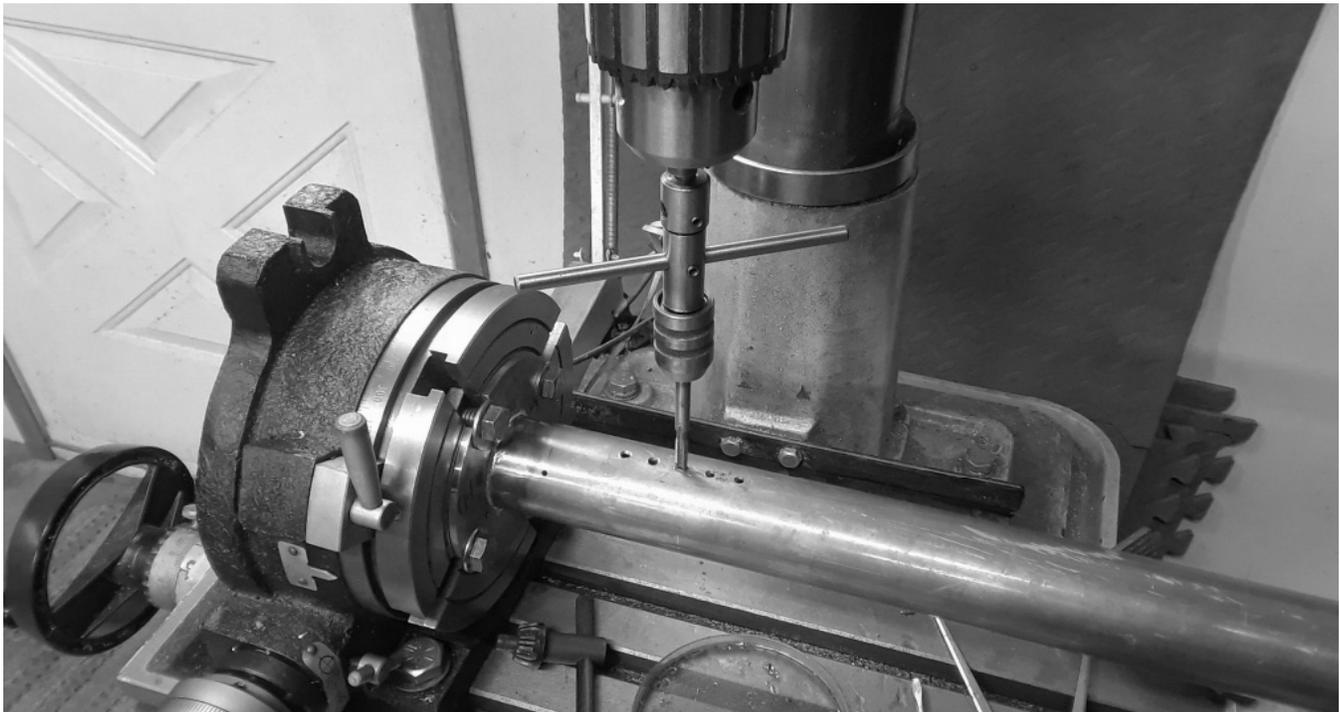
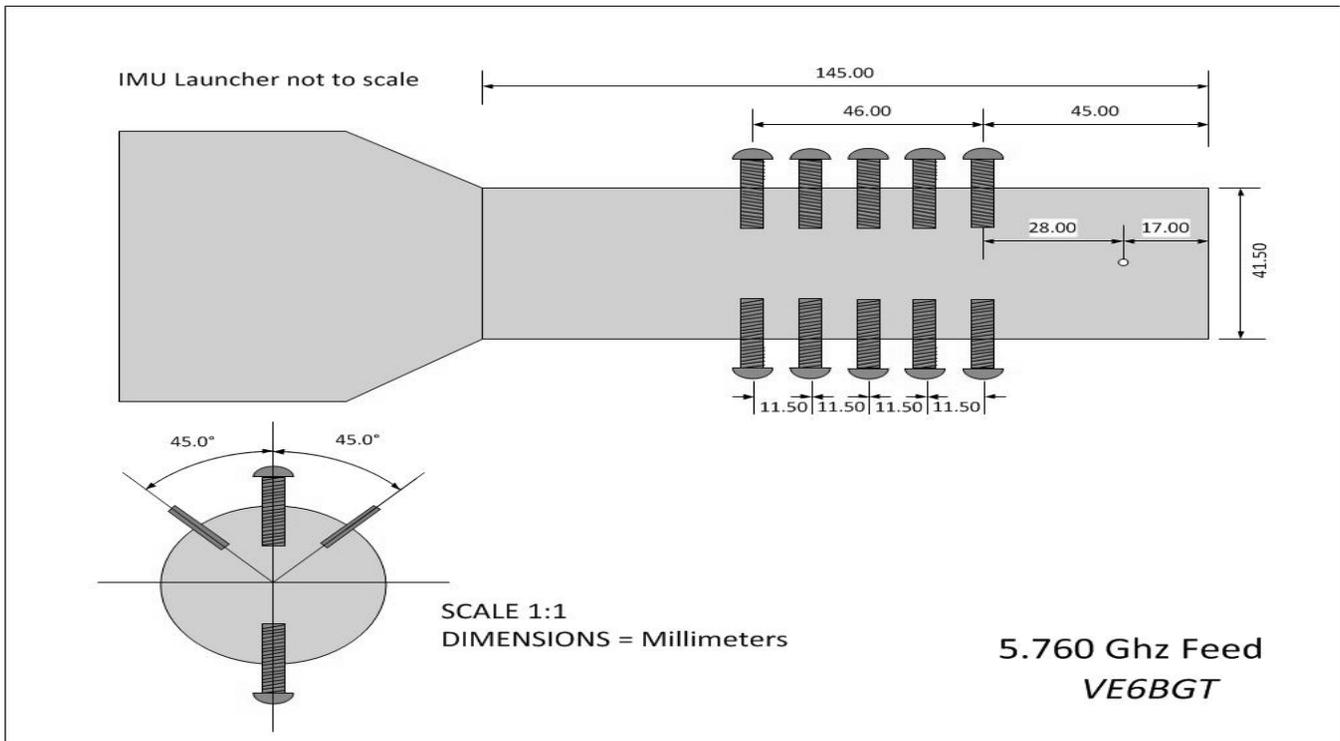
Twisting the IMU launcher section mounted on the end of the feed tube really shifted the RL dip of the TX probe off frequency. This showed me that the IMU section wasn't built concentric or true accurately enough.

The Circular Polarization Test

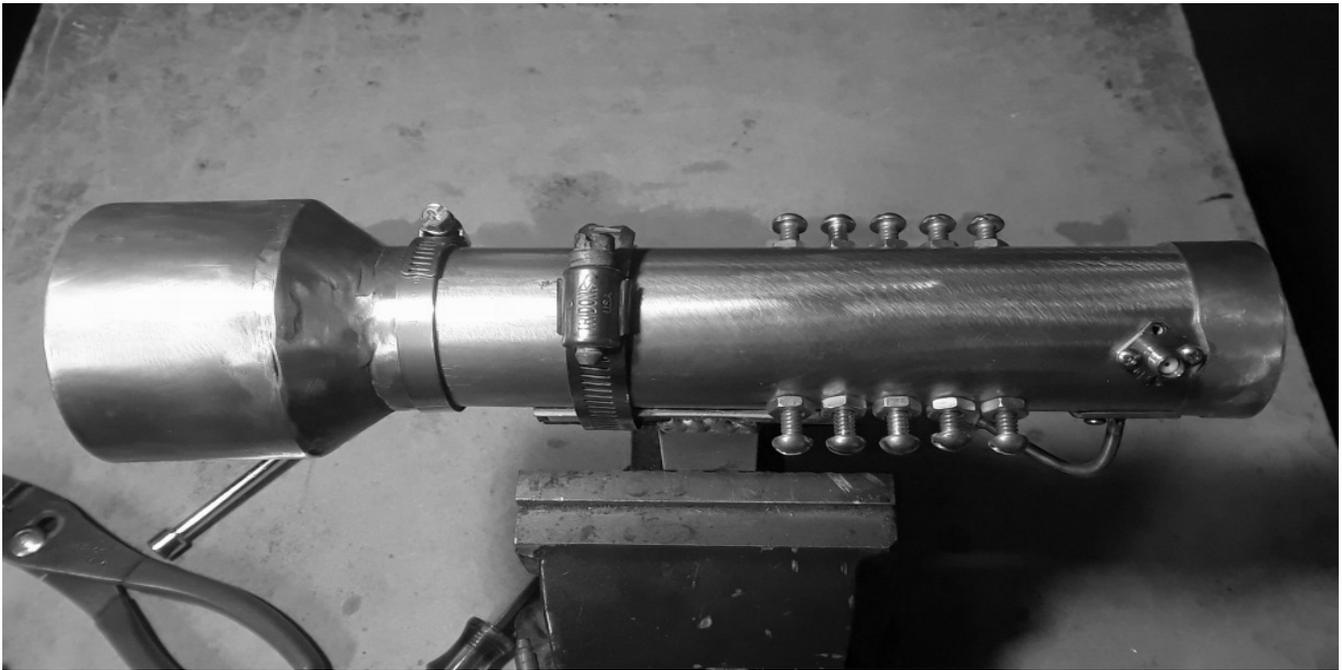


I setup a test to check the circular polarization using a single probe linear feed connected to the sweep generator in CW mode locked by a synchronizer. This linear feed transmits into the squeezed feed under test and the RX probe hooked to a spectrum analyzer for signal detection. I then rotated the linear test feed back and forth and noted the change in amplitude as I went through the different orientation positions. Unfortunately the circular polar (CP) test came up weak, I saw a change in amplitude between 4 to 5 dBm as I rotated the linear test feed. This squeezed feed cannot be adjusted very easily.

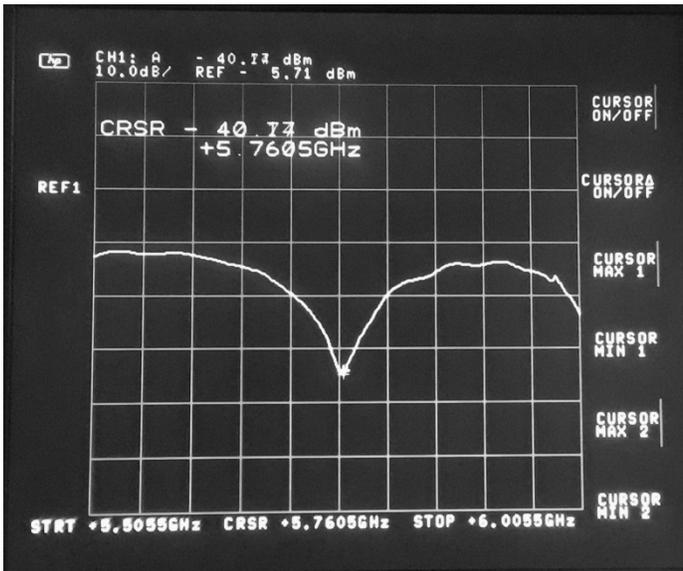
The Next New Idea, The Short Bolt Style Feed



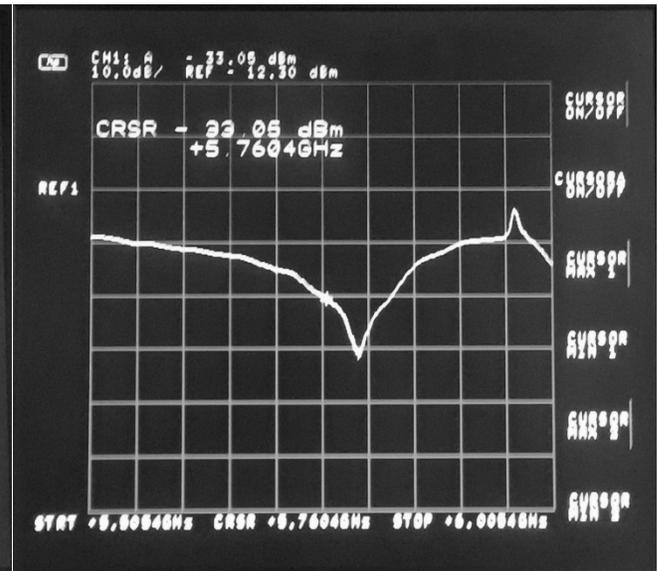
So having good success with the “VE4MA” style of feeds on the lower frequencies, I took the drawings and scaled them to the frequency of 5.760 Ghz. This style of feed uses 10 bolts for achieving the circular polarization and makes it possible to adjust for best results. So back to the milling machine and rotary table to get all the holes accurately drilled and tapped plus the probe connector holes.



The New Constructed Short Feed, TX Probe Using a SMA Connector



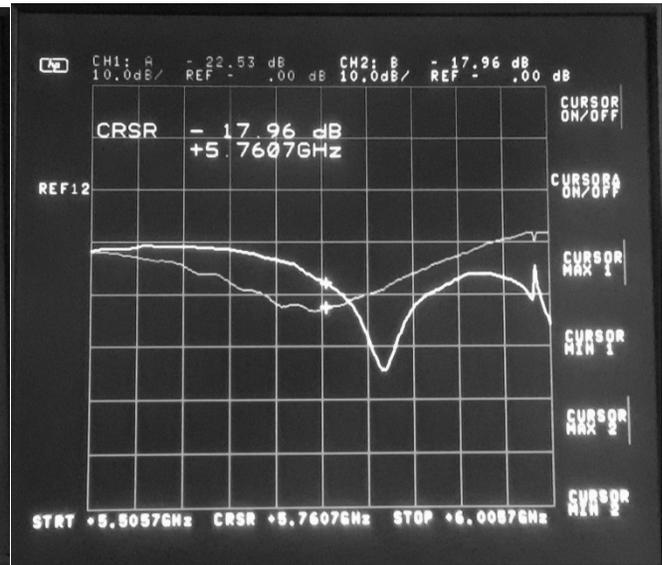
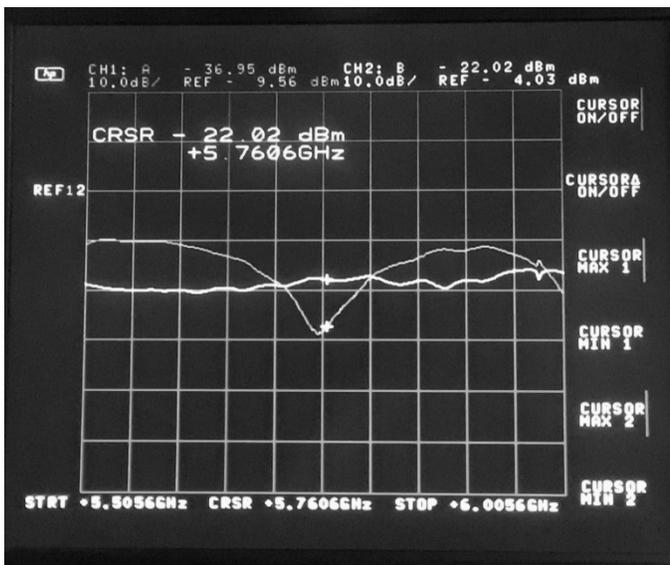
TX Probe, RL = 35 dB



RX Probe, RL = 21 dB

The first frequency test sweeps and probe adjustments turned out not to bad for the first attempt. This was done also by adjusting the back plate and sliding the IMU launcher section back and forth for the best results.

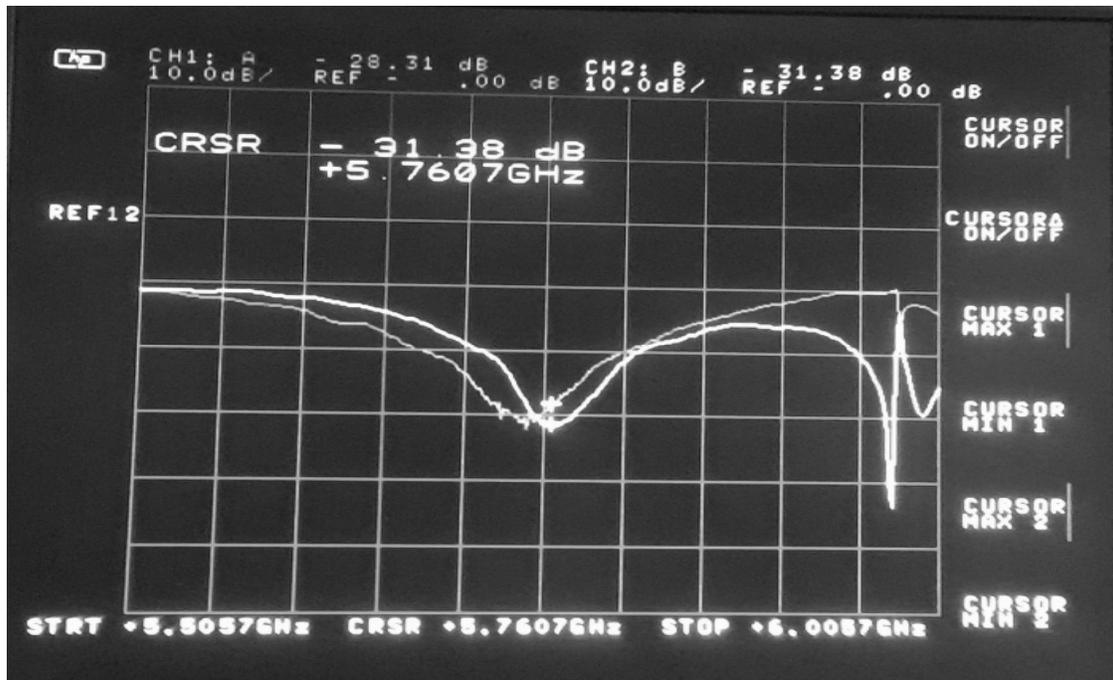
BUT when the isolation was checked between the receive and transmit probes, it was nothing special...



Isolation between probes, around 18 dB

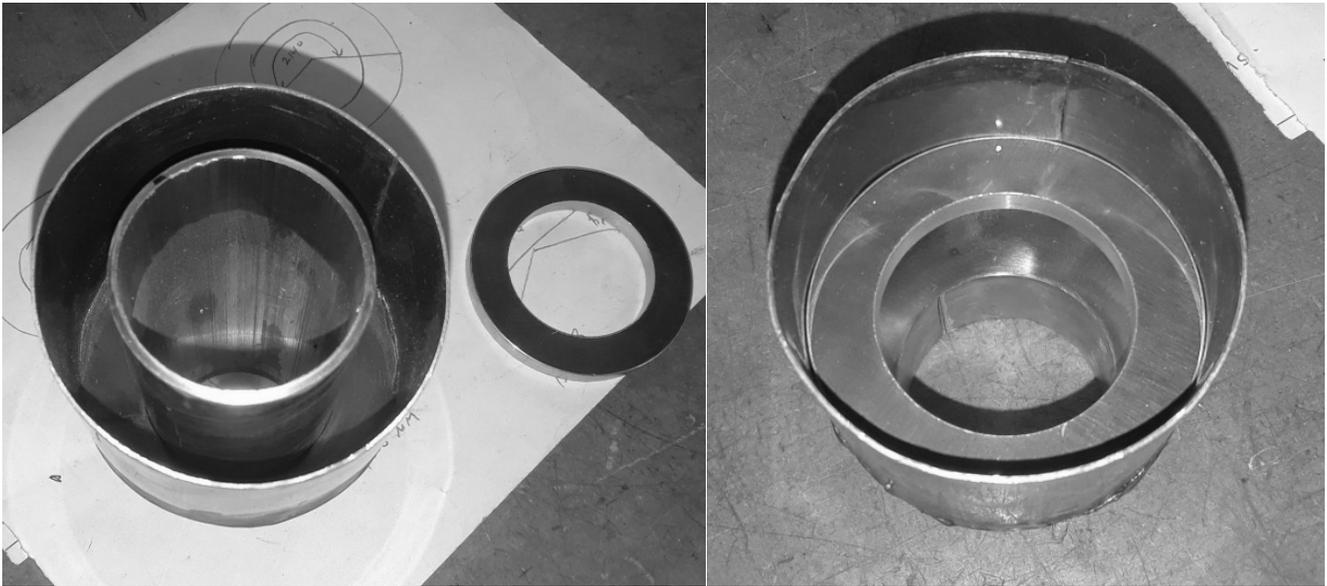
Changed to N connector, ISO 18dB, RL 22dB

I began using a SMA connector and again changed it to an N type as before with the squeezed feed. It wasn't until I set it up for doing the Circular Polarization adjustment that things really came together.

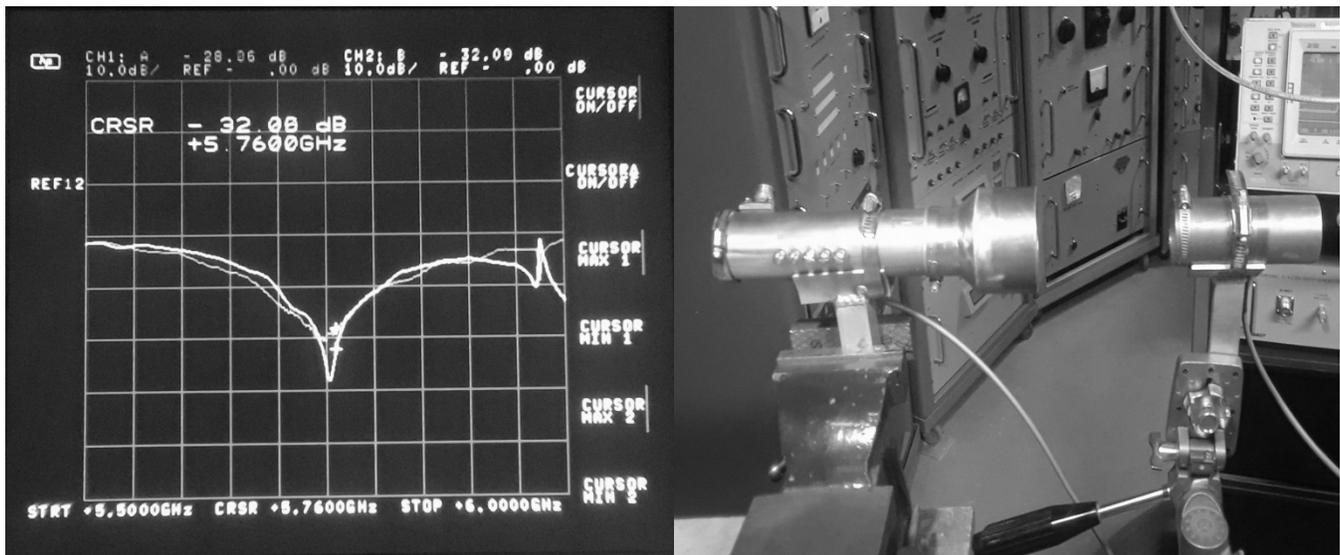


To my surprise after tuning the ten bolts for a real good circular polar test and then going back to recheck the return loss and isolation between probes, the isolation had increased significantly and also nicely lined up the dips on frequency. Readjusting the IMU and back plate and tweaking the bolts even made it better. I then rechecked the circular polarization and saw very little change in it. Once again though I noticed rotating the IMU launcher around the tube shifted the Return Loss dip off frequency. So it was time for a new IMU section.

The New IMU Launcher Construction



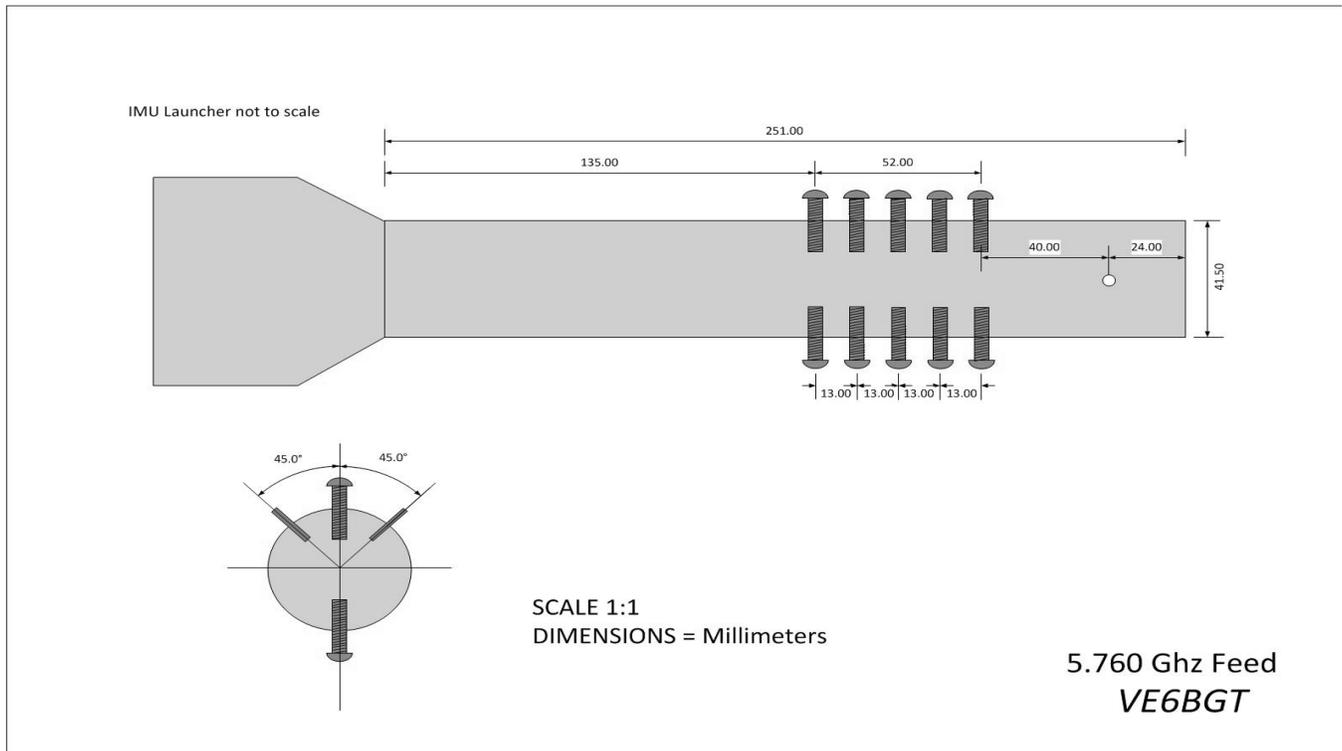
With a special ring lathed from aluminum and placed inside the launcher section, it held the funnel centered and clamped on the main copper pipe waveguide. It was then soldered being held in place making it concentric or true.



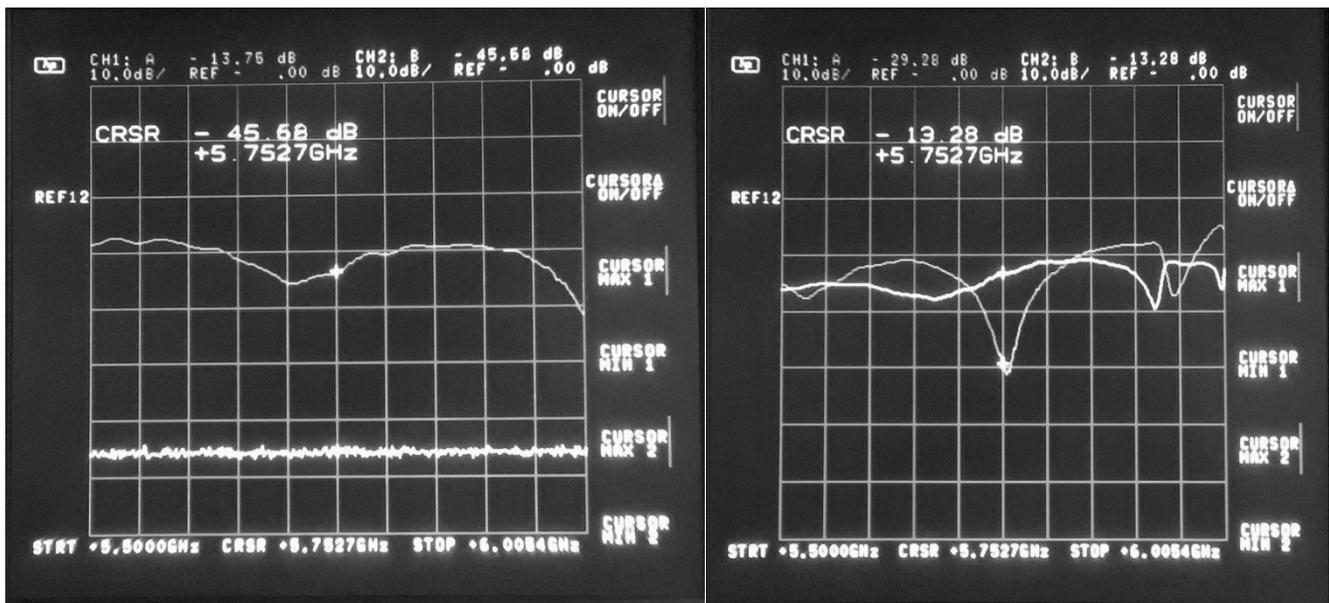
With the improved IMU launcher section mounted the whole feed was once again rechecked and tuned for best results. The TX probe Return Loss was about 28 dB and the Isolation test between the two probes came in around 32 dB. Very little changed now when the IMU was rotated around the feed pipe.

The Circular Polarization was again adjusted and checked; the difference when the test linear feed rotated now was less than 1 dB change in amplitude, the best CP so far.

The Other Big Idea, Would Longer Be Better?

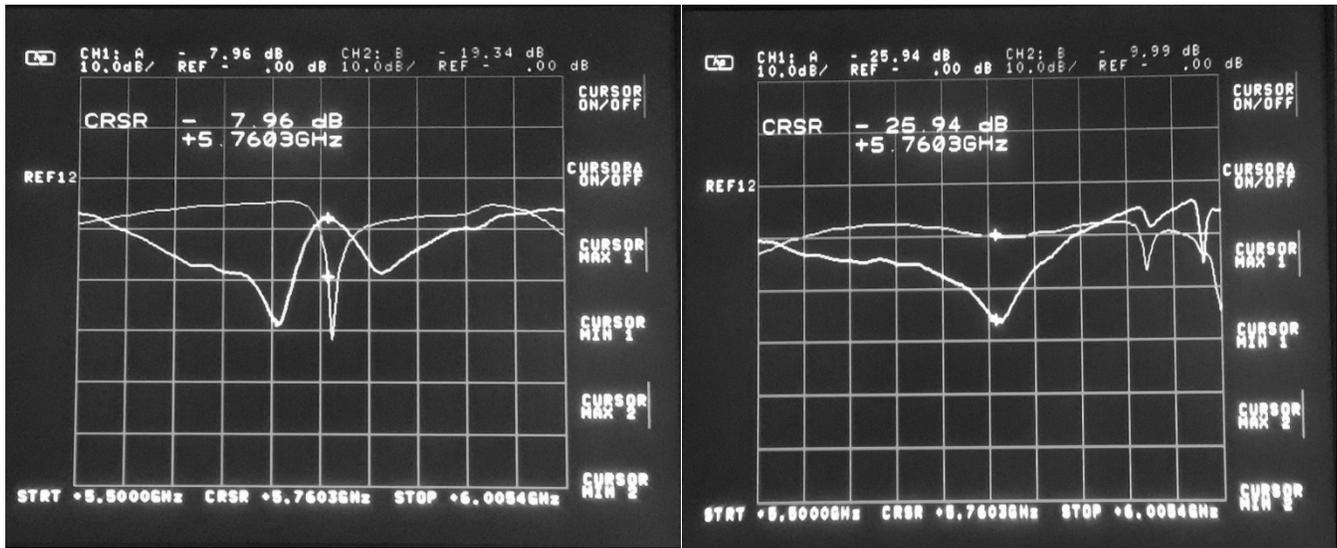


So I wanted to know if I made the overall feed longer and have more room for the CP bolt lock nuts, what would the results be? I started again using a SMA connector on the TX probe.

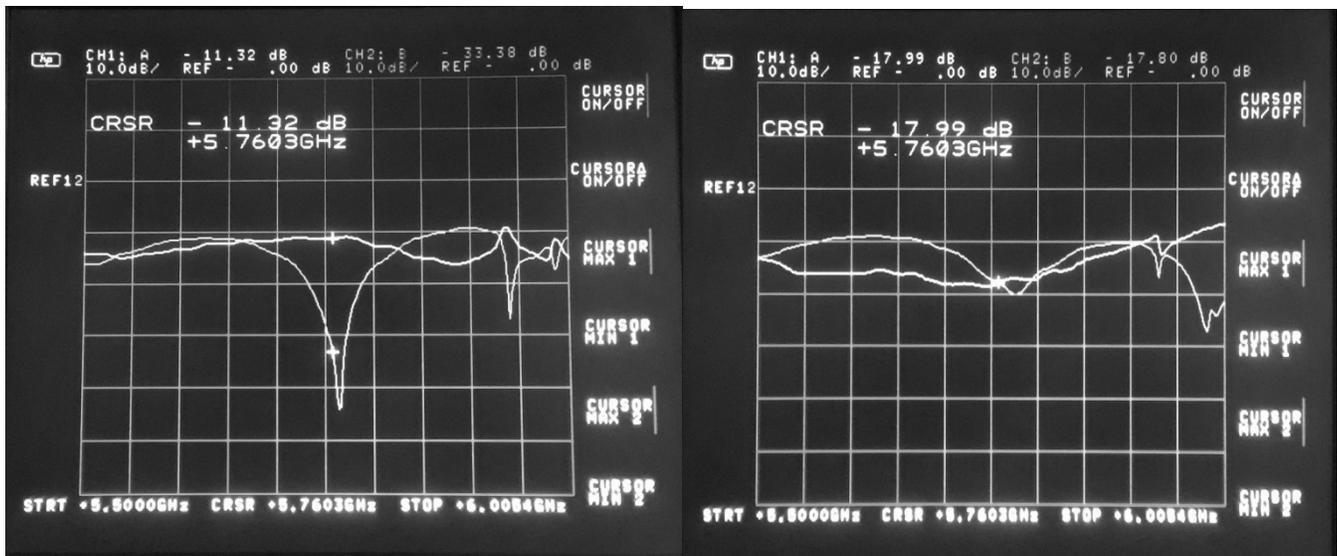


The left pic is trying to adjust the probe only using different thickness of wire. Then on the right pic the IMU launcher and back plate being adjusted for a Return Loss of 29 dB. Isolation again between receive and transmit probes is poor at only around 13 dB. Next was to try the CP bolt adjustment.

The Circular Polarization Adjustment Battle

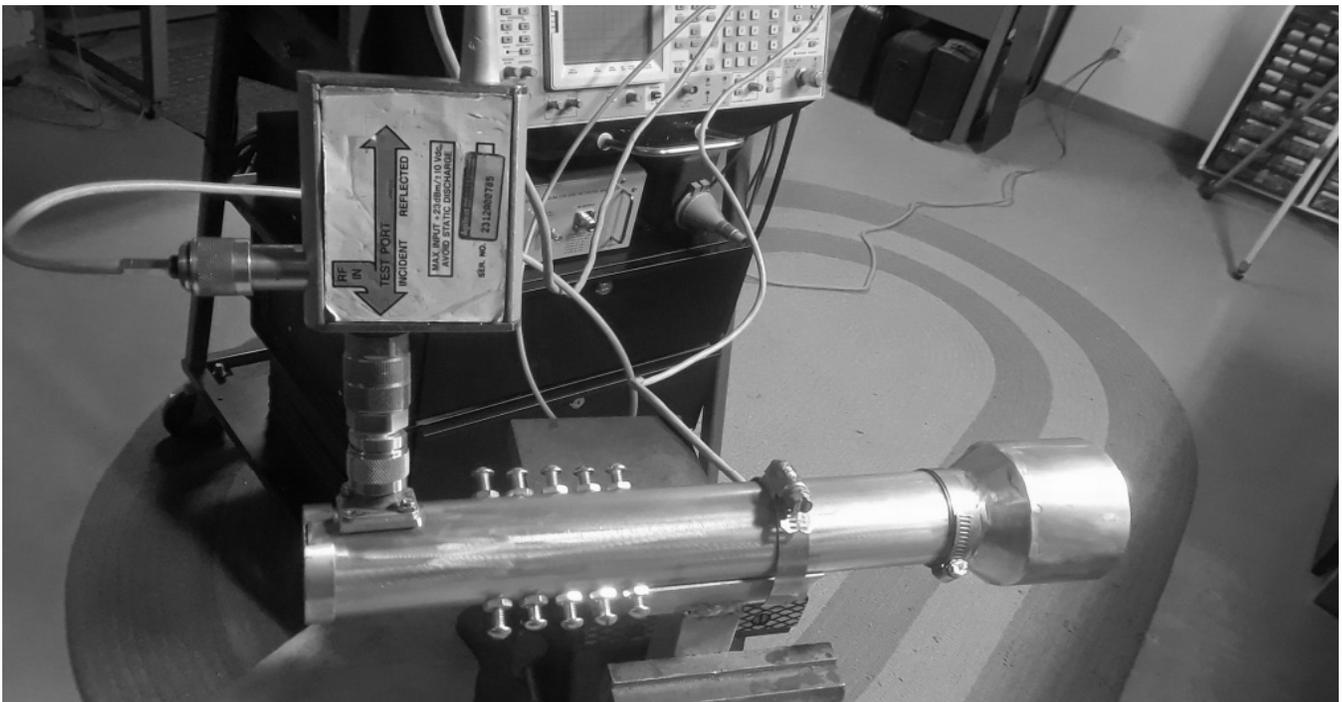


The Multiple Sweep Results Trying to Get the Conditions Right

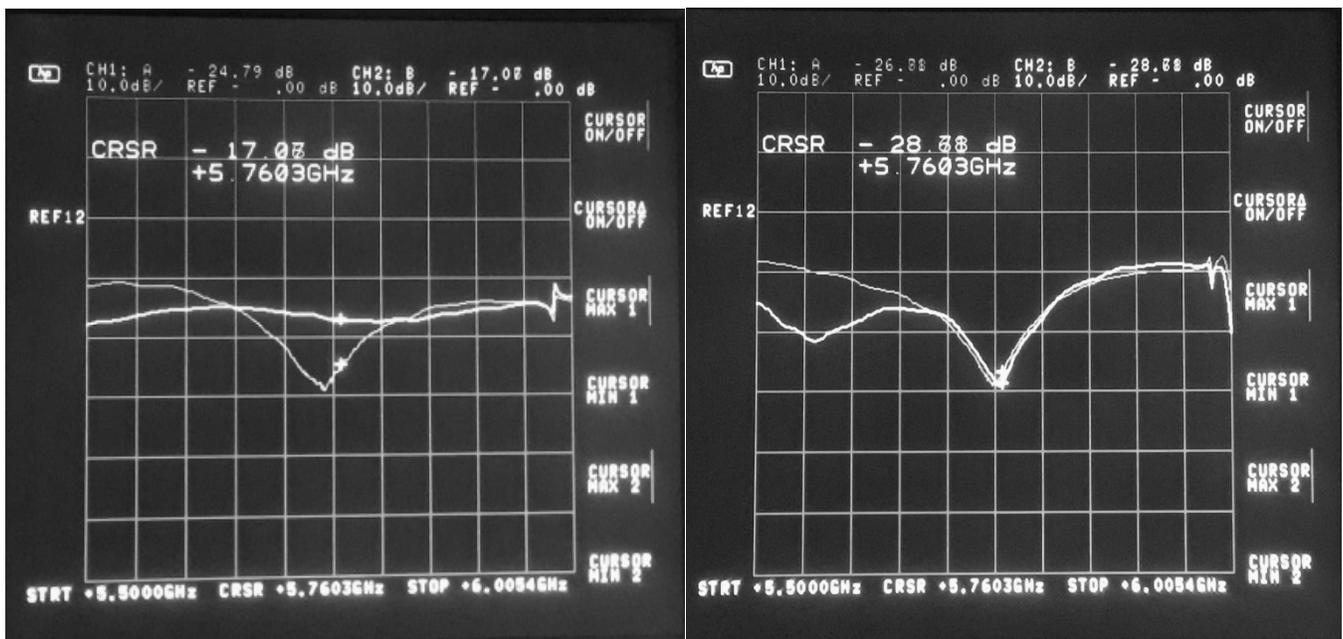


Either the Return Loss Would Be Good, or The Isolation, Never Together

Like before with the shorter feed that tuned well with the CP bolt adjustments, I set out to tune this longer one to improve the Isolation and Return Loss. It was not to be; no matter what I did or tried I could not get both situations to improve together. It was a battle and I spent a long time on it trying to see what would be the sweet spot. The only difference again was that I had a SMA connector on the TX probe, and the spacing between the CP bolts is wider compared to the shorter feed.



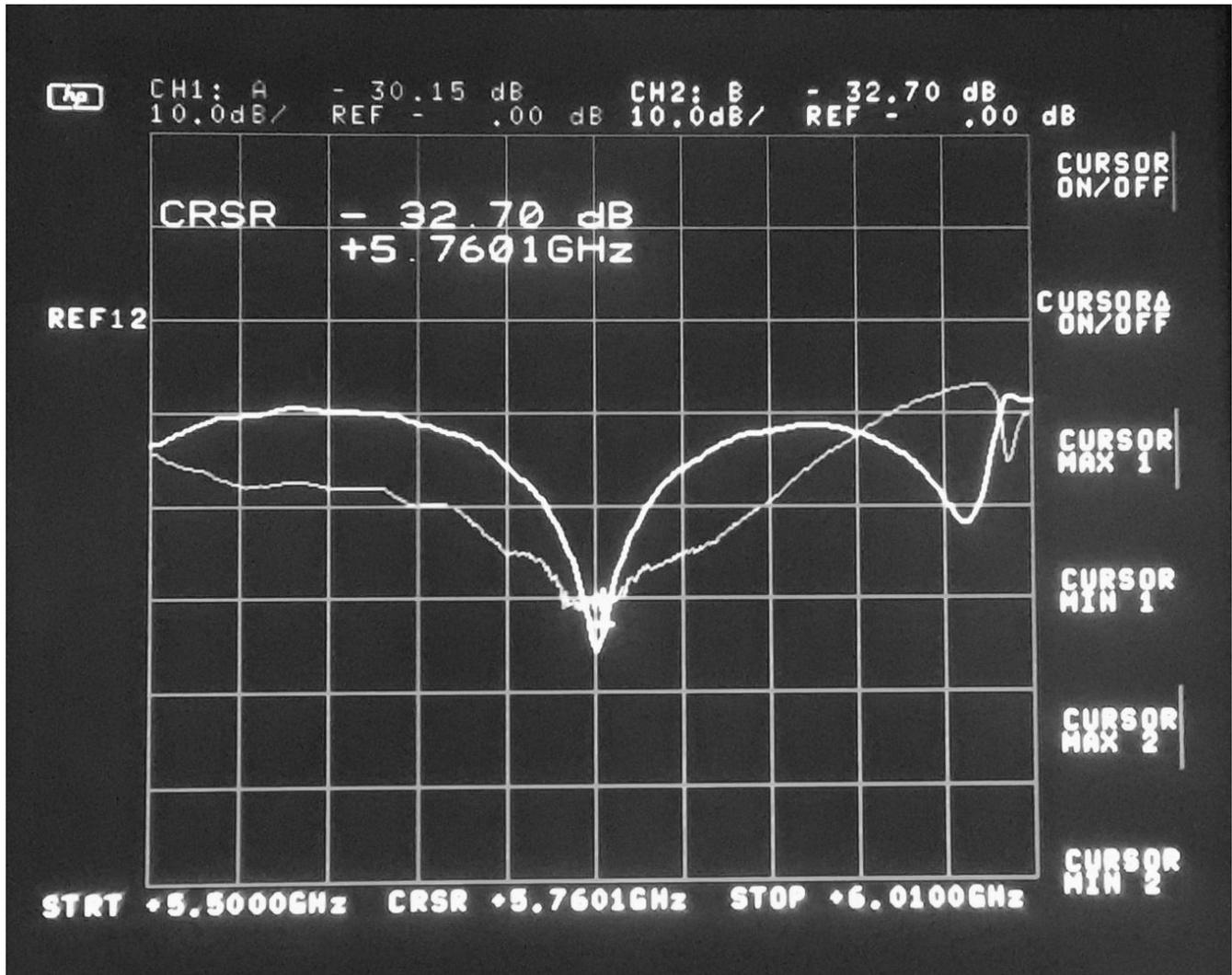
So again I replaced the TX probe SMA connector to an N connector and began adjusting and testing.



Once again it seemed the N connector was the answer. The pix on the left are the Return Loss and Isolation Sweep adjustment before the CP bolts were screwed in at all. The RL is around 24 dB and the Isolation at around only 17 dB, similar to the other feeds at this point.

The right pix shows how good it improved after the ten brass bolts were adjusted for the best Circular Polarization. As before with the shorter feed having the CP adjusted, it greatly improved the RL and especially the Isolation test between probes. RL is now at around 26 dB with the Isolation around 28 dB.

1 - Final Comparison – “Squeezed” Feed



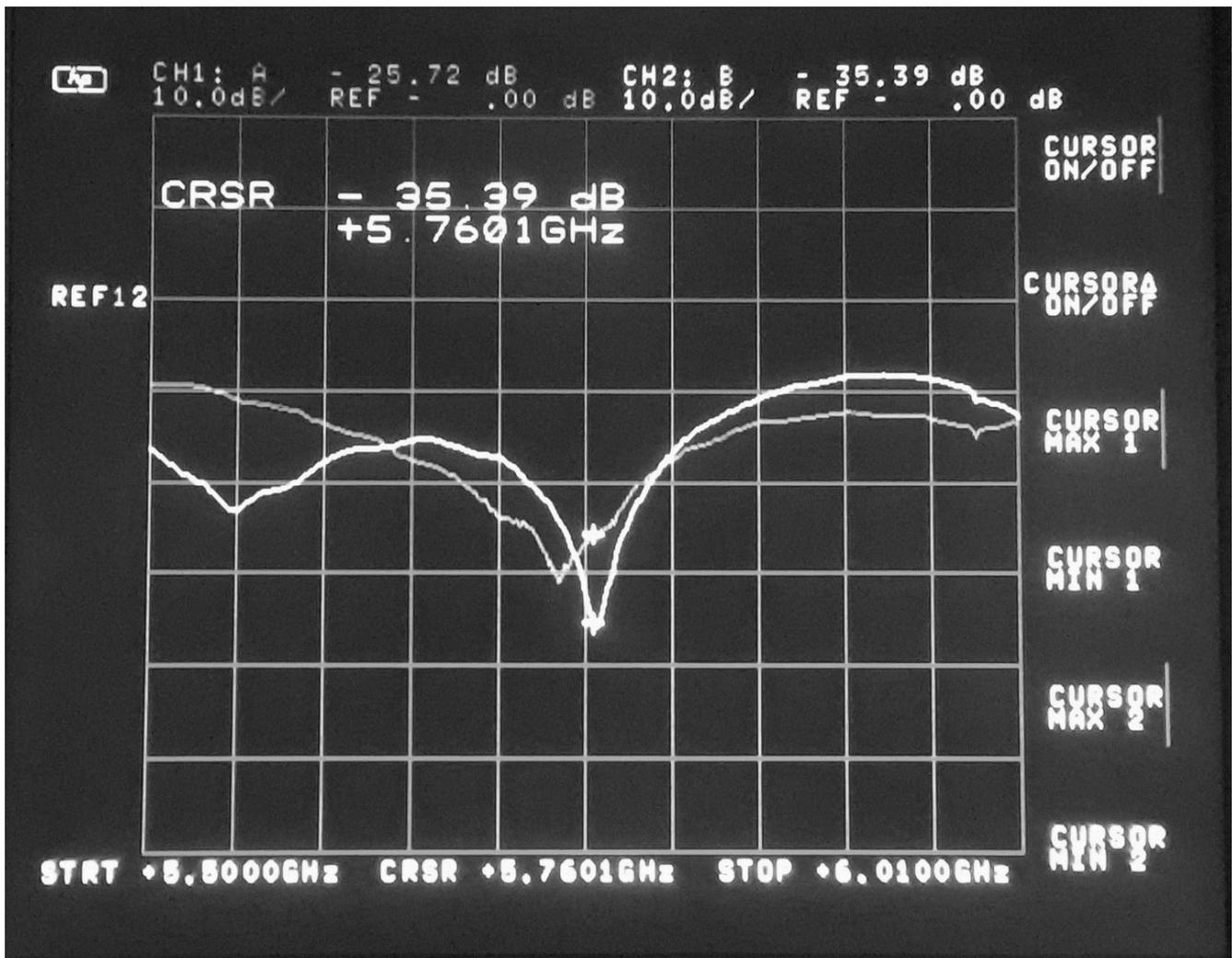
So the final results are in. This final comparison between all three tubes shows that they are very similar but each is better in certain areas.

This pix of the final sweep of the “squeezed” feed shows that the Return Loss of the TX probe measuring around 30 dB, very useable.

The Isolation test between the probes also was good at around 32 dB.

The only weak part of this feed was the Circular Polarization results. When the linear transmitting test feed was rotated there was a significant change in amplitude from the RX probe of the Squeezed feed. As much as 4 to 5 dBm in amplitude was seen on the spectrum analyzer when the test signal was rotated. Since this type of feed has fixed plates it is hard to impossible to adjust for a better Circular Polarization result.

2 – Final Comparison – Long Bolt Polarizer Feed



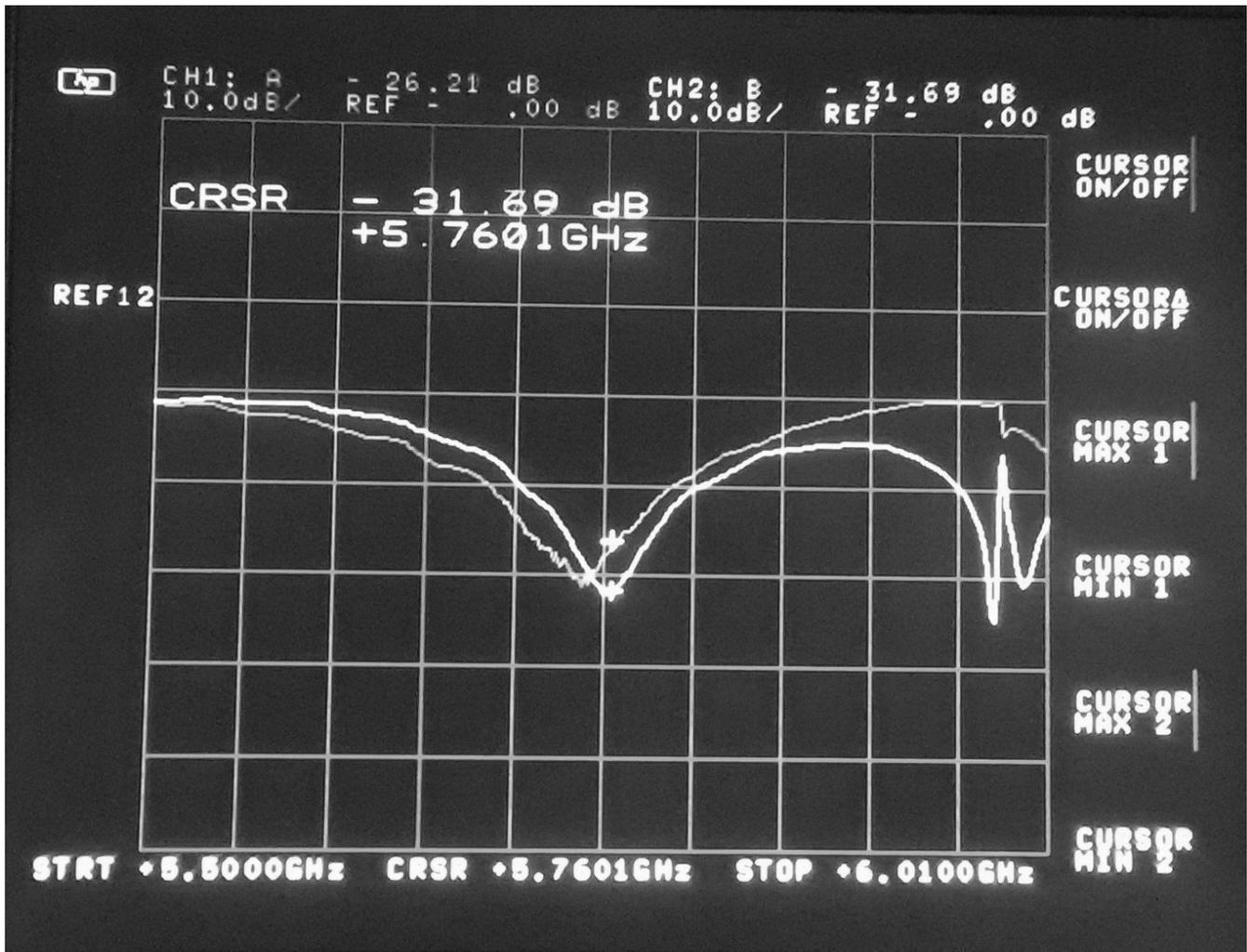
This feed with the longer copper pipe waveguide and using the ten brass bolts for the Circular Polarization was at first a battle to get adjusted and tuned.

The Return Loss came in at around 25 dB and the Isolation between probes finally at around 35 dB.

The Circular Polarization test came in better than the “Squeezed” feed. The amplitude difference I saw was around 2 dBm only. This could be because of the wider bolt pattern of the ten brass CP adjustment bolts.

Again this is quite a usable feed.

3 – Final Comparison – Shorter Bolt Polarizer Feed, The Winner



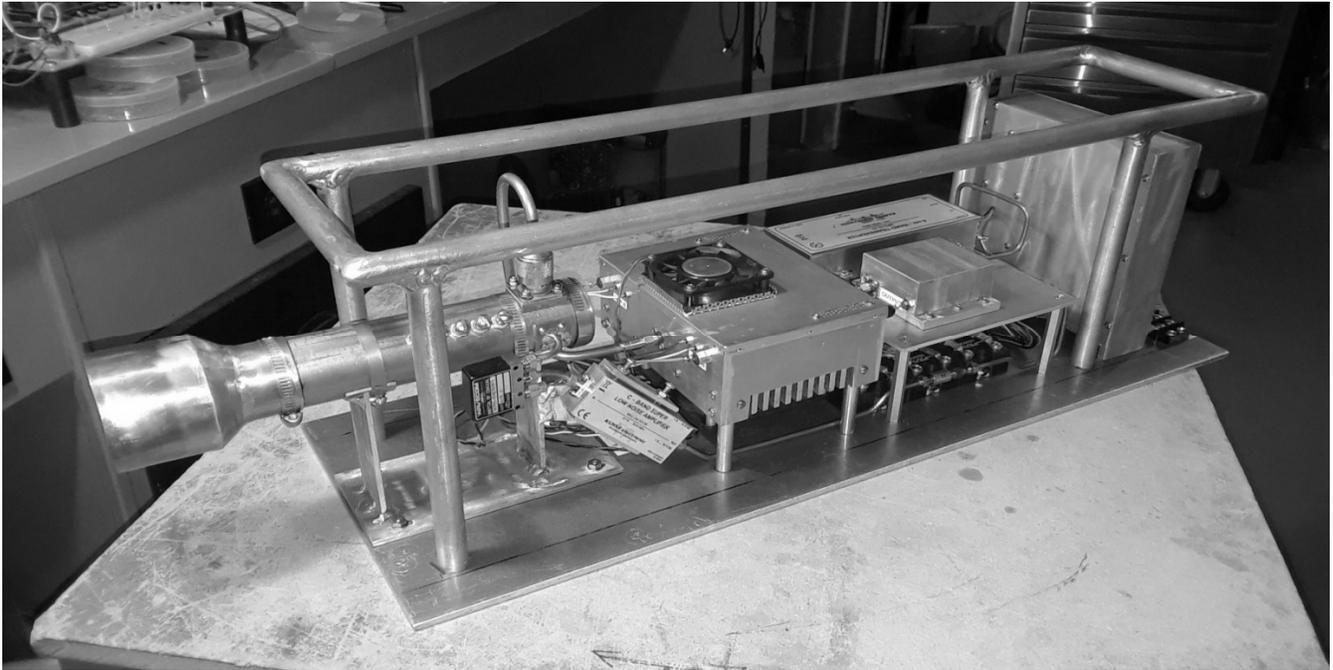
Again the final test of this shorter copper pipe wave guide type feed is excellent. It did seem to tune up easier than the longer one, the Return Loss around 26 dB with the Isolation between probes at 31 dB.

The biggest improvement was with the Circular Polarization test. The difference I measured when rotating the test linear feed was less than 1 dBm at the most. This could possibly result from the ten brass bolts being closer together than the wider spaced ones on the longer feed pipe.

Other Final Conclusions or Results

It seemed that using N connectors made getting better results a lot easier for all three feeds. I was going to use SMA originally but remembered with the 9 CM feed I built before, N connectors just worked better. I did try different types and makes of SMA connectors with the same results. I wouldn't write off SMA connectors all together though. The other big thing I noticed was the effect the IMU launcher had on tuning. To exchange the IMU launcher to scalar rings to see if a solar noise test improved, this would require that the whole feed to be retuned or adjusted.

The Complete 6 CM Feed Assembly



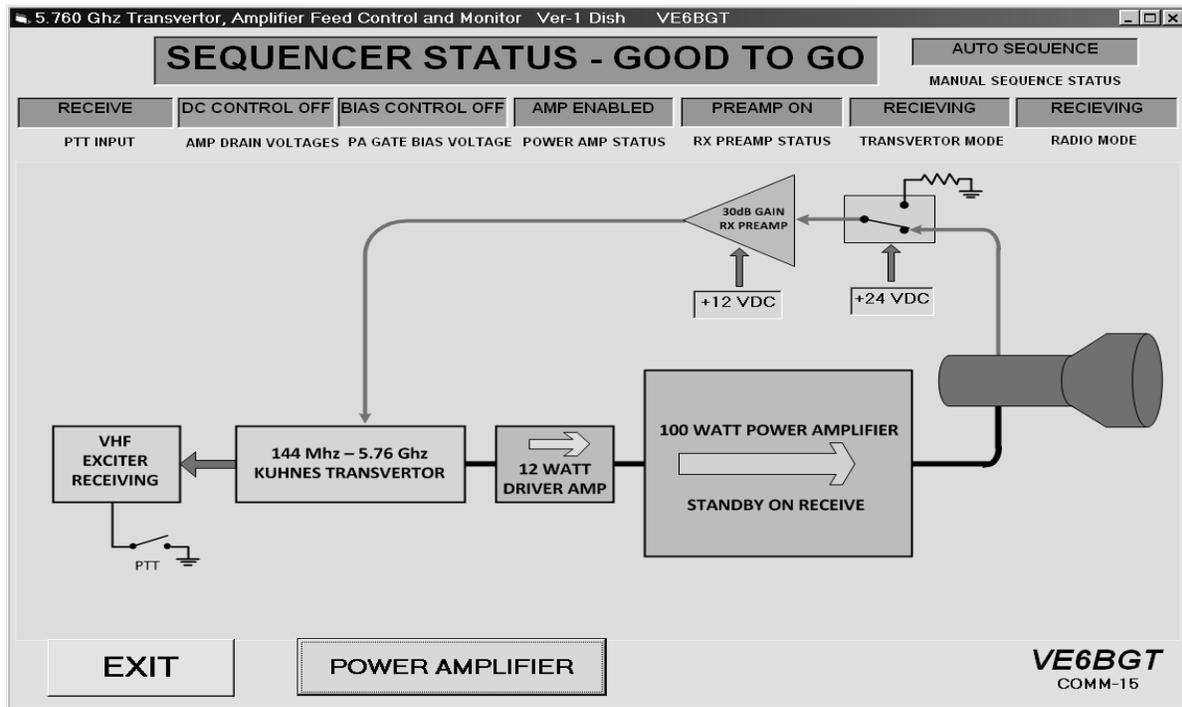
This is a pix of the new complete 6 CM feed assembly ready to be placed on the dish feed mounting shelf. It's controlled and monitored by a PIC microprocessor in the back box, and uses a Kuhnes transvertor and receive preamp. Lots of testing of the circuits was done to get to this point of the project.



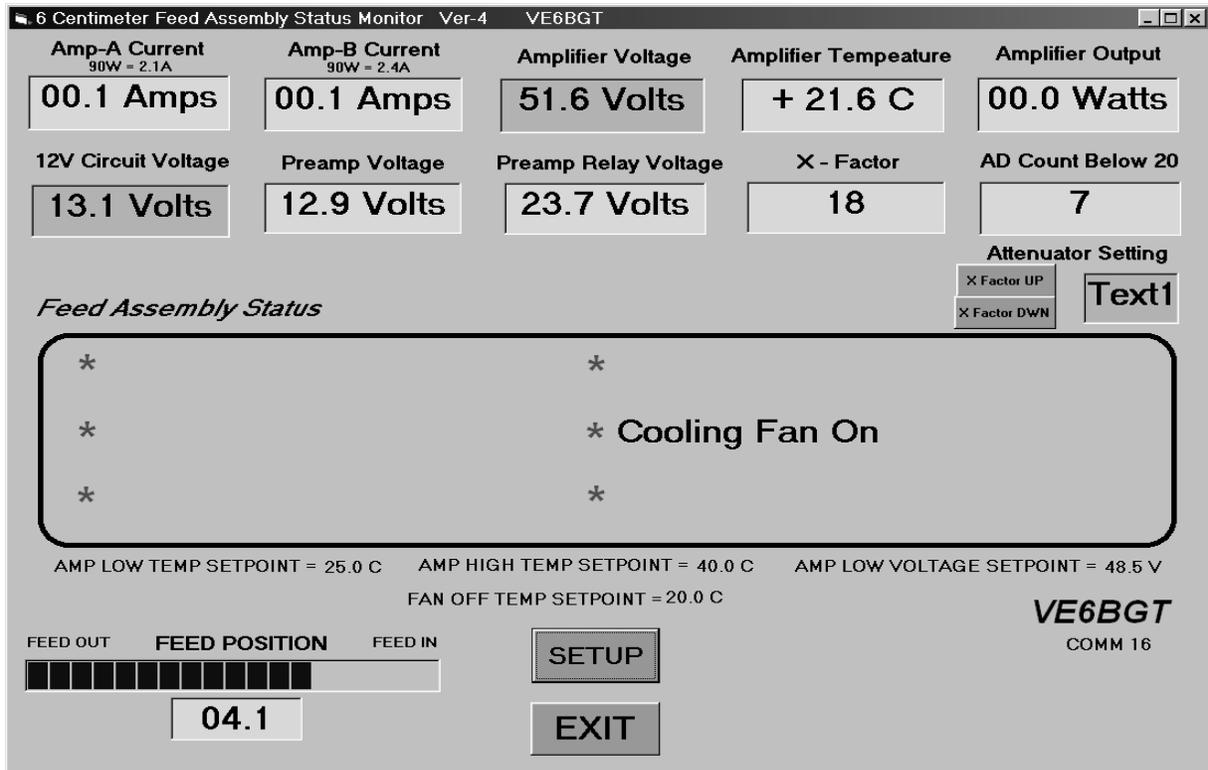
The new 6 CM feed assy. mounted on the 21 foot dish for final testing. Underneath is a feed position actuator to slide the assembly in and out for peaking on solar noise etc.

This actuator has a position sensor that is connected to the PIC circuit and position data is sent back to the shack for monitoring.

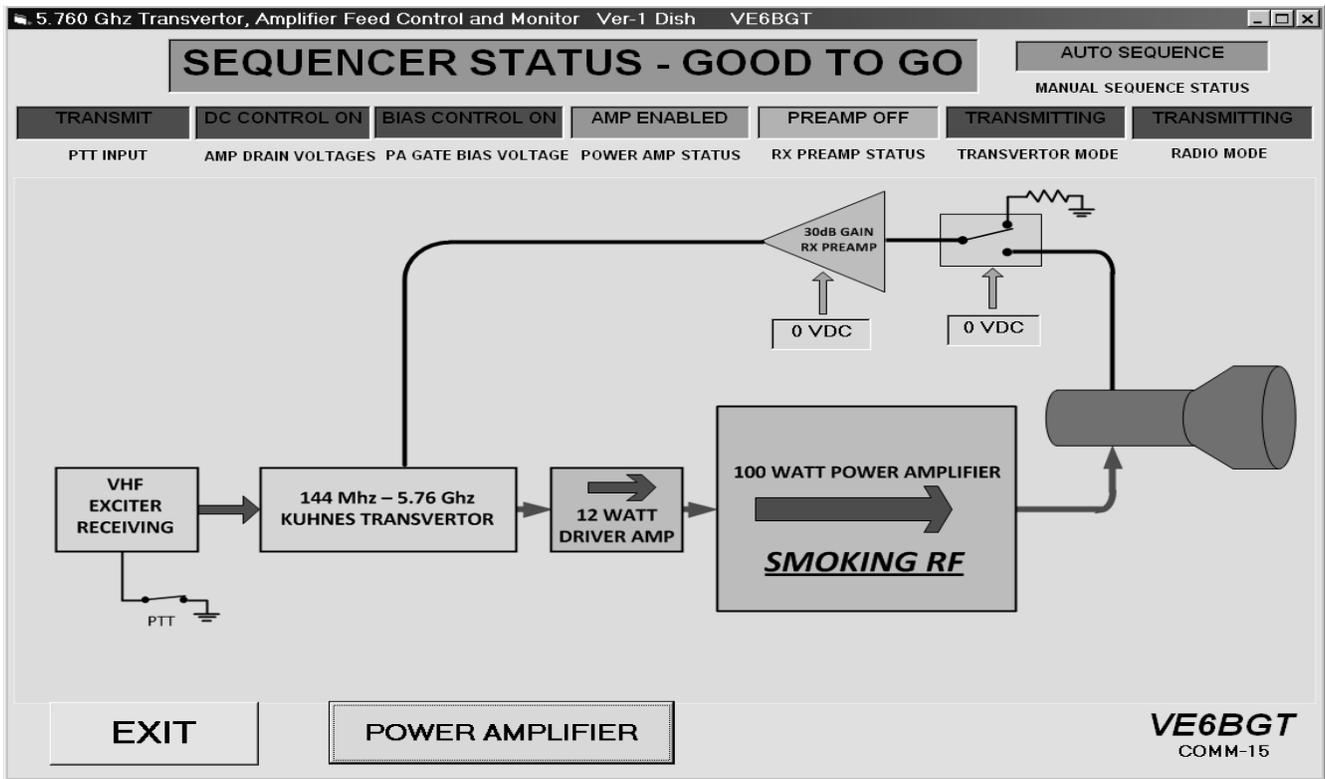
The 6 CM Monitoring and Control Screens



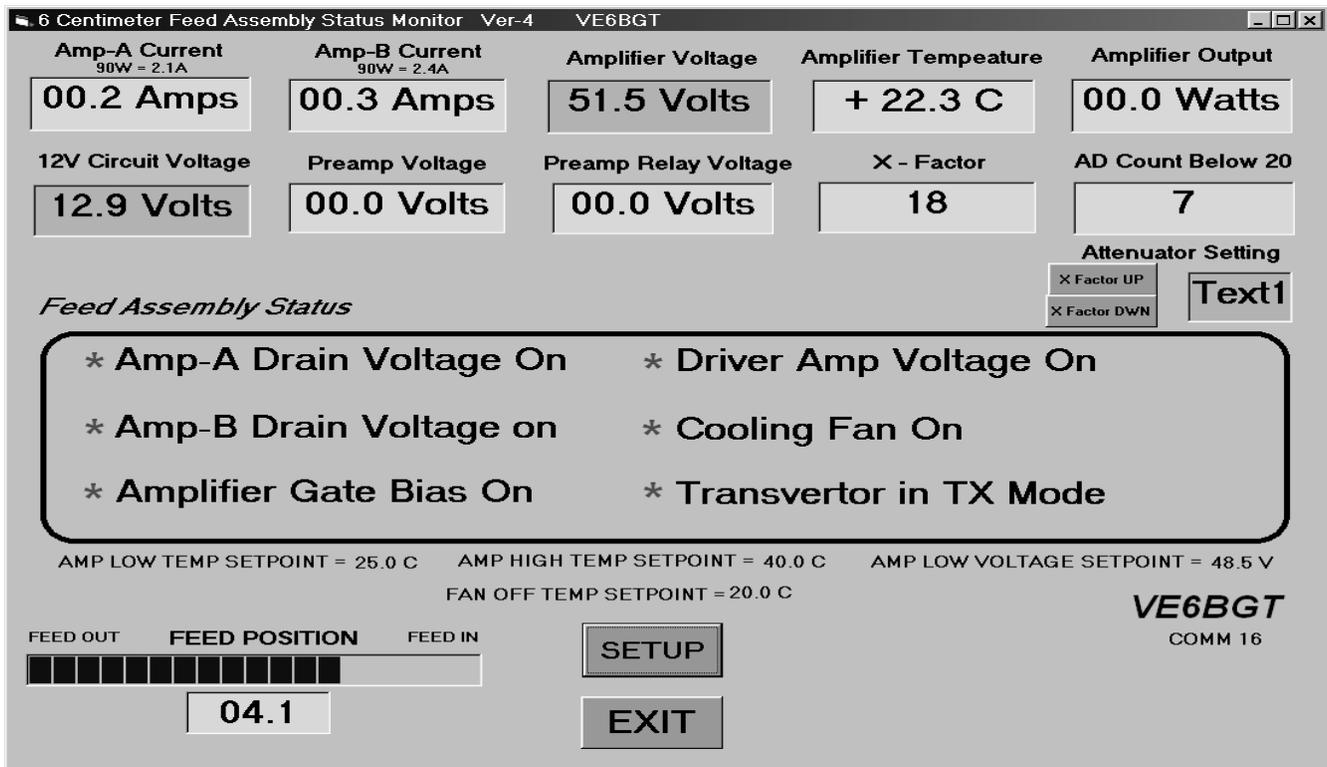
A screen shot of the 6 CM system status in receive mode.



The 100 watt amplifier and various voltage point status monitoring in receive mode.



6 CM System Status in Transmit Mode.



Amplifier Circuit and Voltage Points in Transmit Mode, Keyed Only.

IMPEDANCE MATCHING TECHNIQUES FOR RF & MICROWAVE CIRCUIT DESIGN

Tom Apel – K5TRA

OVERVIEW

The need to match impedance is unavoidable by anyone working in RF. Power transistors are not 50 Ω input and optimum loads at the transistor output are not 50 Ω . Optimum noise match impedance at a LNA FET input is not 50 Ω . This paper presents techniques to first, design lumped LC circuits for impedance matching. Next, a straightforward technique to convert a lumped design into a printed circuit realizable form is presented.

Bandwidth needs for amateur radio are relatively low; perhaps 10% or less. Techniques to broadband match over an octave (or more) use synthesis techniques. For those readers interested in broadband synthesis, please read my paper on that subject:

<http://k5tra.net/TechFiles/Broadband.pdf> .

Q·BW LIMITS TO MATCH PERFORMANCE

The Q of the load sets natural limits to the achievable impedance match. The best possible match reflection coefficient is limited by the product of load Q and BW, the % bandwidth. This

is expressed mathematically as: $\rho \geq e^{\left(\frac{-\pi}{Q \cdot BW}\right)}$.

In terms of this reflection coefficient, SWR is expressed as: $SWR = (1 + \rho)/(1 - \rho)$.

This best case match performance is represented graphically in Fig.1. The curves indicate ideal performance limits not practically achievable; but serve as a useful bound for expectations. That said, a 1.05:1 SWR ideal limiting case could easily correspond to an SWR of 1.1:1 or more in a practical finite order circuit. So, for a 10% BW match, the load Q shouldn't be greater than 8. A load Q greater than 8 will require a match bandwidth less than 10% for the same match performance.

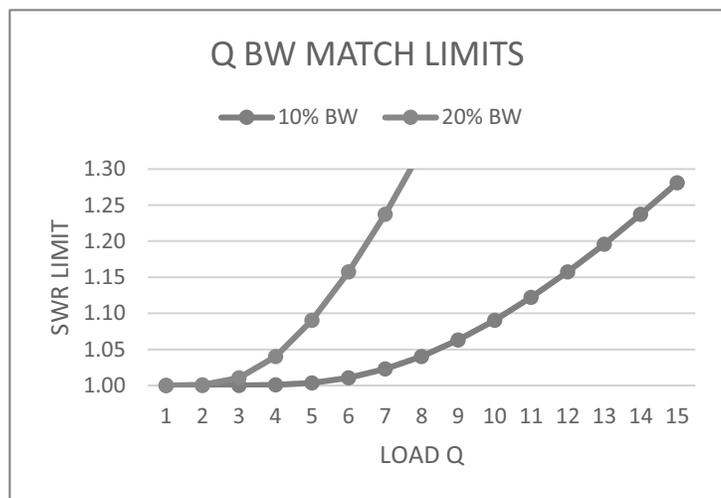


Figure 1 Ideal limit to match

TWO ASPECTS OF IMPEDANCE MATCHING

Transformers and impedance transforming networks are an important part of impedance matching. It would be unusually fortunate to need a match to a purely resistive load. Usually, the load impedance is complex; so, transformer action and reactance (or susceptance) absorption is required from the circuit. In most narrow band cases, impedance matching can

be thought of as resonating the reactance (or susceptance) followed by the necessary transformer action to match the source resistance (often 50Ω).

The primary approach to the design of lumped element matching circuits will be based on the Smith Chart. For completeness, I would like to first like to briefly discuss use of several non-LC impedance transformers in matching networks. If the reactive part of the load impedance is resonated with an opposite series reactance, then the remaining task is to simply cascade an appropriate transformer. The same is true if the load is viewed as a parallel RC or RL, where an opposite susceptance is added to resonate the load and an appropriate transformer is cascaded. Several obvious transformers are quarter-wave transmission lines or short coupled-line transmission line structures such as Ruthroff or Guanella transformers. For a thorough presentation of these topics, please read my paper on that subject: <http://k5tra.net/TechFiles/Transformers.pdf> .

THE SMITH CHART

The easiest way to design narrow band impedance matching circuits is with the Smith Chart. It is an extremely useful graphical tool that unfortunately is viewed as “mysterious” to the uninitiated. A bit of explanation will hopefully remove the mystery. First, consider Fig. 2. The reflection coefficient, ρ , is the ratio of reflected to forward voltage at the load. It is a complex number representable as either (real, imaginary) or (magnitude, angle). The maximum magnitude of reflection with passive loads is 1.0 (total reflection). This is plotted as the unit circle in Fig. 2 and represents the outer boundary of the standard smith chart.

REFLECTION COEFFICIENT: ρ

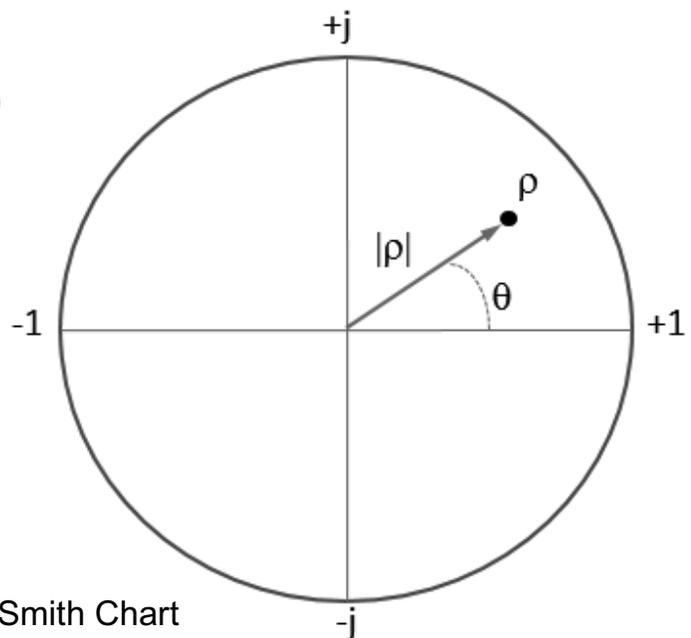
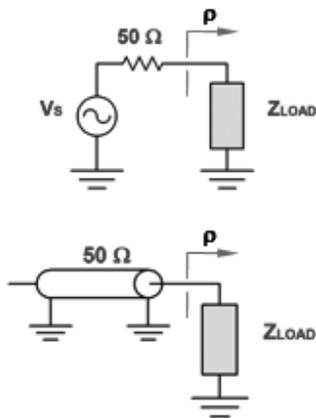


Figure 2 Origin of the Smith Chart

In other words, ALL possible passive reflection coefficients are contained inside this circular region. Equivalently, ALL possible passive load impedances are representable by points contained inside this circular region.

Next, consider the two representations of all possible impedances with a constant real part of 50Ω , illustrated in Fig. 3. In order to represent all loads with the same real part (50Ω in this case), the rectangular chart contains a vertical line infinitely long. Clearly, this page isn't large enough to capture all of that line. In contrast, that constant 50Ω line is fully captured within the Smith Chart on the right side of Fig. 3

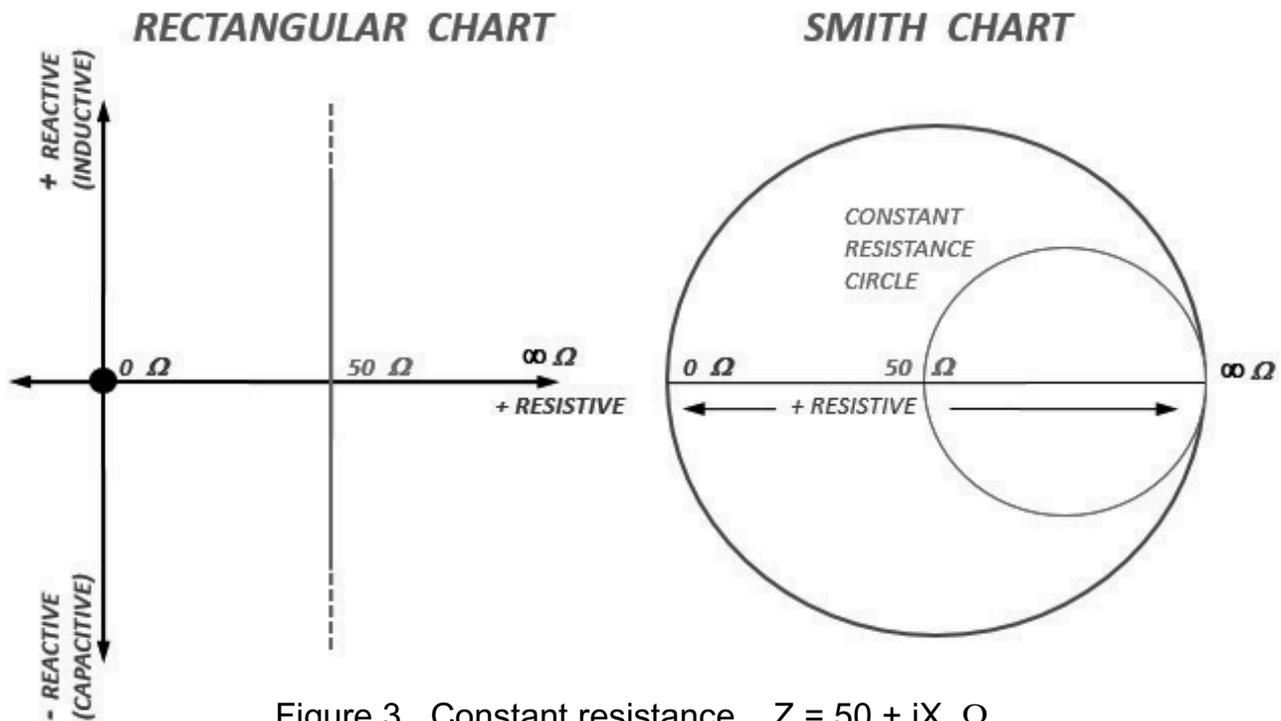


Figure 3 Constant resistance $Z = 50 + jX \Omega$

Impedance is a complex number with a real part and an imaginary part: $Z = R + jX$. Impedance represents a series connection of a resistor and a reactance. A positive reactance is inductive and a negative reactance is capacitive. Constant real lines (resistances) become circles on the Smith Chart. Note also that the positive real axis from the rectangular chart maps into the horizontal line that divides the Smith Chart into upper and lower halves.

If we consider constant reactance lines, either $+j 50 \Omega$ inductive or $-j 50 \Omega$ capacitive, the rectangular chart contains an infinitely long horizontal line. This is shown in Fig. 4. The upper half of the Smith Chart represents all possible positive real, inductive impedances. Similarly, the lower half of the Smith Chart represents all possible positive real, capacitive impedances.

When more constant R and constant X lines are plotted in Smith Chart form, the impedance chart results. This is illustrated in Fig. 5.

It should again be restated that the impedance chart represents series connections. For example, if we start with a load impedance point on the chart $50 + j 50 \Omega$, and add a series reactance, we will obtain a new location on the same real line. Only the reactance is changed. Tuning with series LC elements stays on the same real circle.

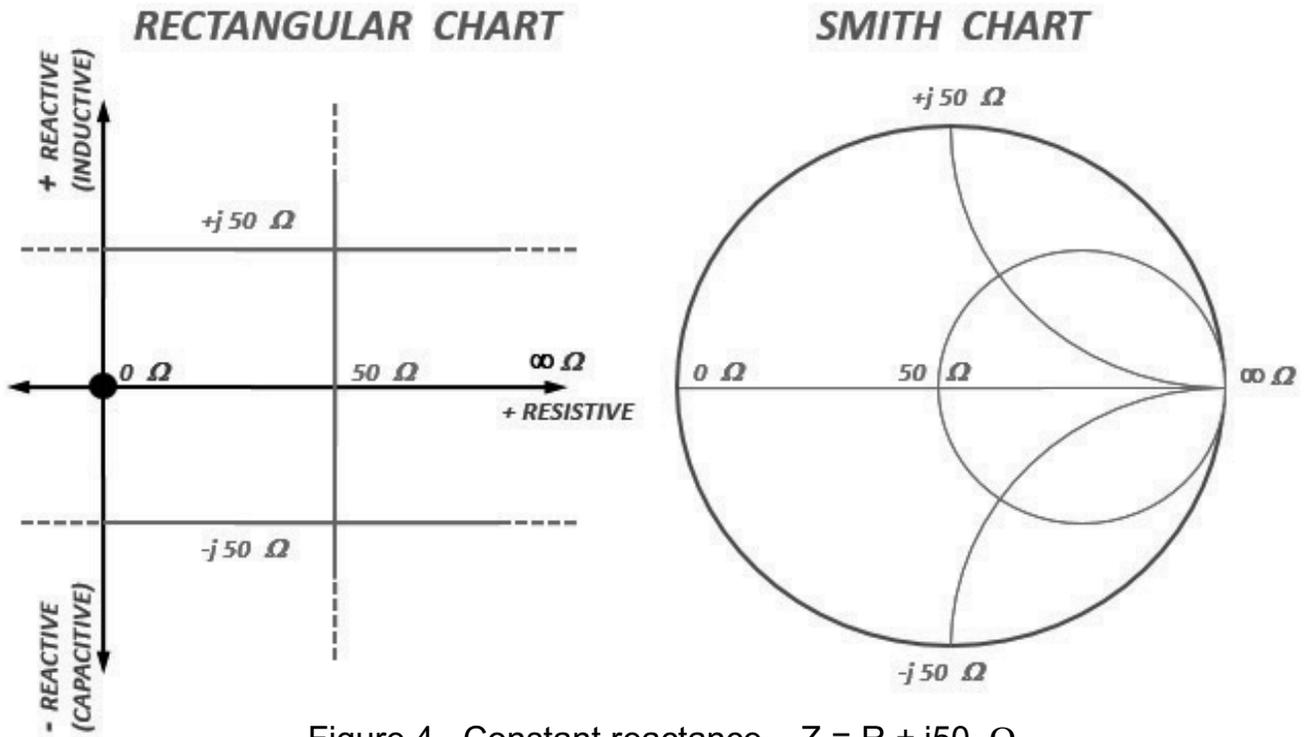


Figure 4 Constant reactance $Z = R \pm j50 \Omega$

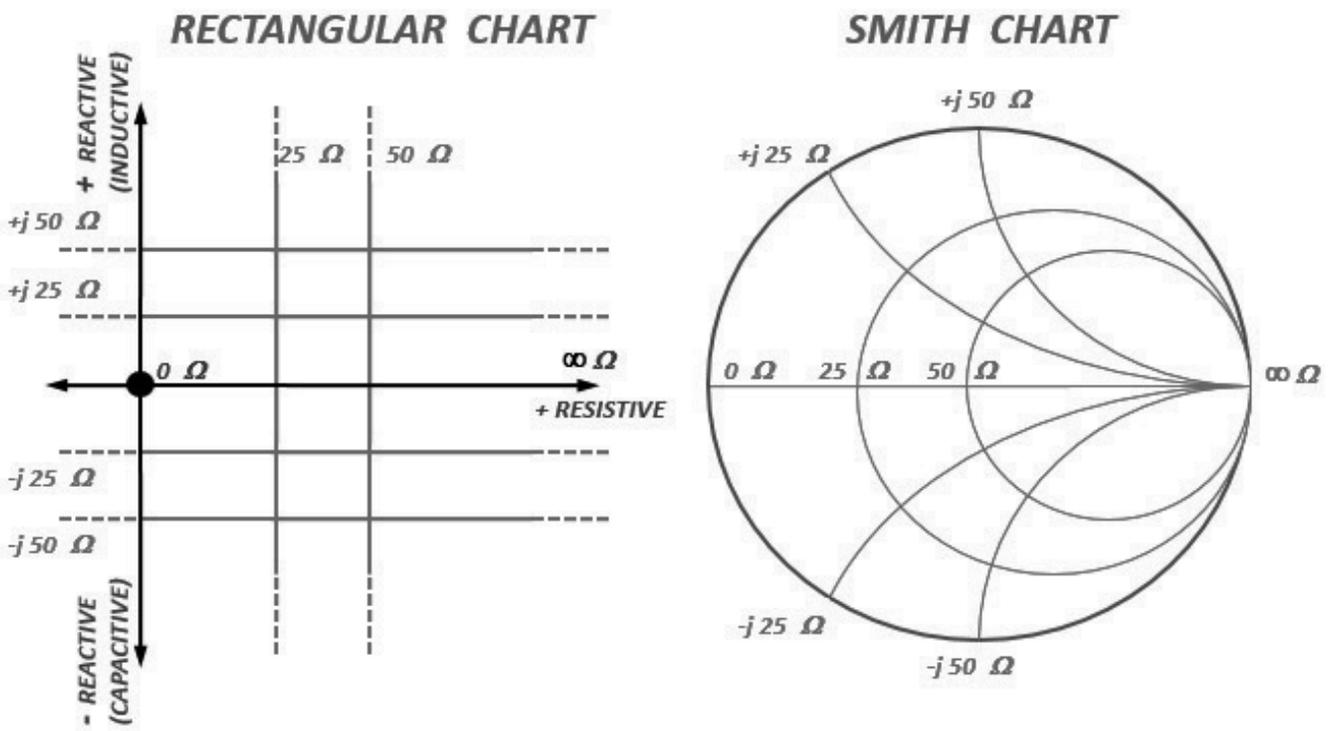


Figure 5 More constant R and X lines produce the impedance view chart

Parallel connections of resistors and capacitors or resistors and inductors are best represented in terms of admittance. Adding or subtracting parallel elements directly adds or subtracts to the admittance. Like impedance, admittance is complex with a real part called conductance (G) and an imaginary part called susceptance (B). Admittance is expressed as: $Y = G + jB$. Units of admittance are $1/\Omega$ or S (Siemens).

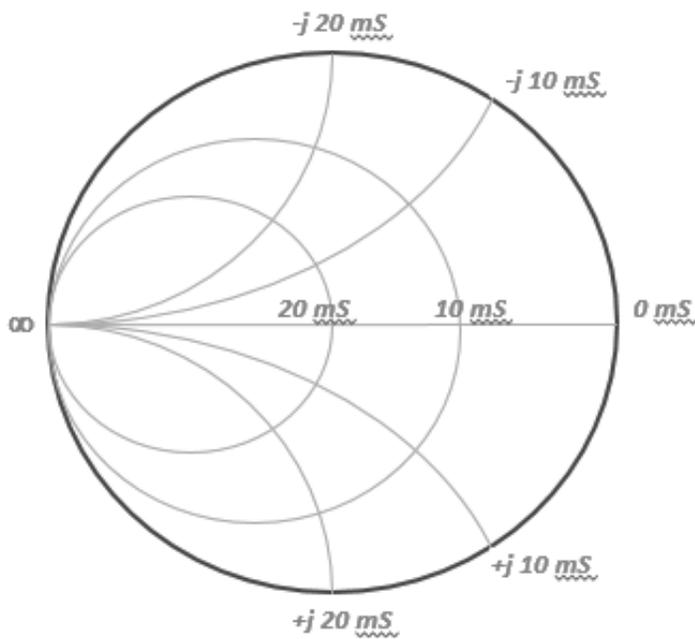


Figure 6 The admittance chart

The admittance chart is shown in Fig. 6. Just as constant resistance contours are circles in the impedance chart, the admittance chart has constant conductance contours as circles. Note that the 20 mS (20 miliSiemens) circle passes through the center of the chart. The center of the chart is $50 \Omega = 1/(20 \text{ mS})$, or alternatively $20 \text{ mS} = 1/(50 \Omega)$. Constant susceptance (imaginary part of admittance) contours are arcs similar to the constant reactance arcs in the impedance chart. The upper half of the chart is inductive (negative sign susceptance) and the lower half is capacitive. This is consistent with the nature of those regions in the impedance chart.

If the impedance and admittance charts are overlaid, a very powerful tool results. This composite chart is sometimes called the immittance chart.

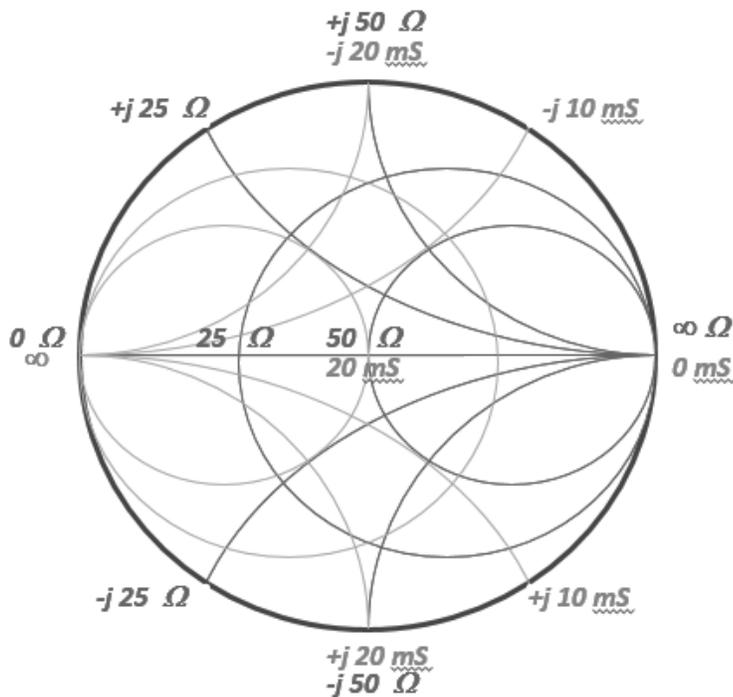


Figure 7 The immittance chart

Each point on this chart can be interpreted as a series resistance and reactance or as its equivalent parallel form. As shown in Fig. 2, each point on the chart can also be interpreted as a reflection coefficient in a (center of chart reference) 50Ω system. Sometimes you will see the chart published in a "normalized" form where the center of the chart is 1 instead of 50Ω or 20 mS . In that case, you will need to multiply all normalized impedance numbers by 50 and all normalized admittance numbers by 0.020.

The power of the overlaid chart is that it allows to one simultaneously switch back and forth between series and

parallel views. This will allow a matching circuit to be built element by element.

As an example of lumped element matching with the immittance chart, start with a 25 Ω load and match it to 50 Ω. In Fig. 8, the 25 Ω point, A, is on the horizontal real axis and on the constant resistance (25 Ω) circle.

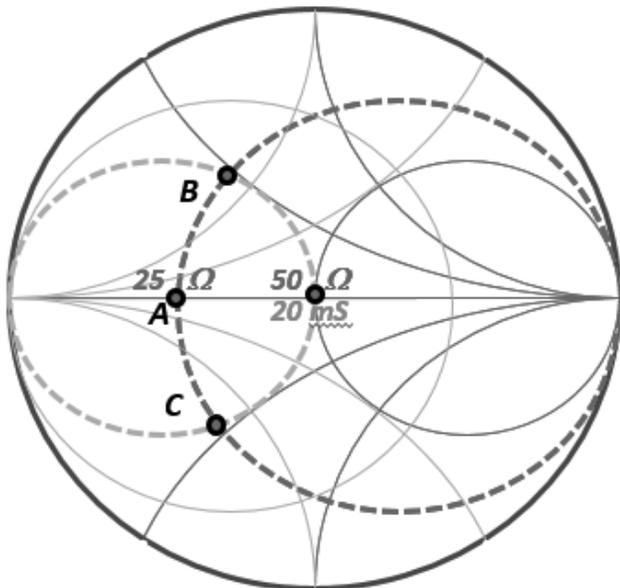


Figure 8 Lumped element matching

As inductive reactance is added, we move upward on the 25 Ω circle. If the correct amount of inductance is added, we will move to point B. Alternatively, capacitive reactance could be added to move from A to C. Note that both A and C are on the 20 mS constant conductance circle. Next we use the parallel representation contours. By adding parallel capacitance, we can move from B, down the 20 mS circle to the center of the chart. We have a match from 25 Ω to 50 Ω with a series inductor followed by a shunt capacitor ! Similarly, a series capacitor followed by a shunt inductor path, moving from A to C to the center, could be used for an alternate matching circuit. Quantitatively, the amount of series reactance and shunt susceptance can be read directly from the chart. The L and C values are the

calculated from the reactance and susceptance values. Series L or C element values are calculated from the reactance shifts as read from the Smith Chart:

$$L_S = \frac{|X_L|}{2\pi F} \quad \text{and} \quad C_S = \frac{1000}{2\pi F |X_C|} \quad , \text{ where units are L(nH), C(pF), X(}\Omega\text{) and F(GHz).}$$

Similarly, the shunt L or C elements are calculated from the susceptance shifts as read from the Smith Chart:

$$C_P = \frac{|B_C|}{2\pi F} \quad \text{and} \quad L_P = \frac{1000}{2\pi F |B_L|} \quad , \text{ where units are L(nH), C(pF), B(mS) and F(GHz).}$$

Examples of larger transformation matches can be seen in Figs. 9 and 10. In both of these examples, two section (4 element) matches are used to match 5 Ω to 50 Ω. Figure 9 does this with two lowpass sections (series L, shunt C, series L, shunt C). Figure 10 does this with a lowpass and a highpass section (series L, shunt C, series C, shunt L) to achieve a bandpass response. In both cases, the first section matches 5 Ω to 15.8 Ω and the second section matches 15.8 Ω to 50 Ω.

In addition to the immittance chart, please note the constant Q (= 1.5) “football” shaped contours. It is important to maintain the low and similar loaded Q in both match sections. This will allow a similar loss to be achieved in each. The lower the loaded Q of a matching section, the lower the loss. Losses in lumped matching circuits are due primarily to series resistance in inductors. Losses in capacitors are typically much lower than in inductors. A useful measure of inductor loss is Qu, the unloaded Q. The insertion loss of a two element section is expressed

mathematically as: $\text{Loss(dB)} = 20 \log_{10} \left[\frac{(Q_U - Q_L)}{Q_U} \right]$. Clearly, high Q_U inductors provide low circuit losses.

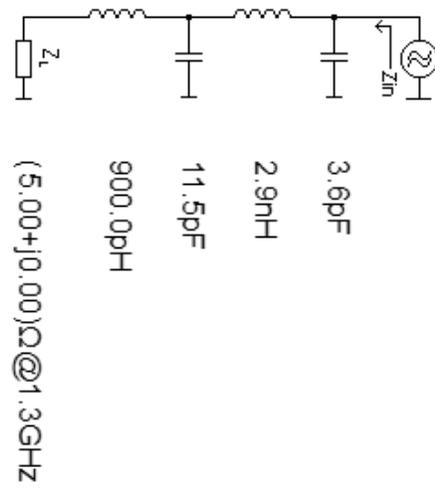
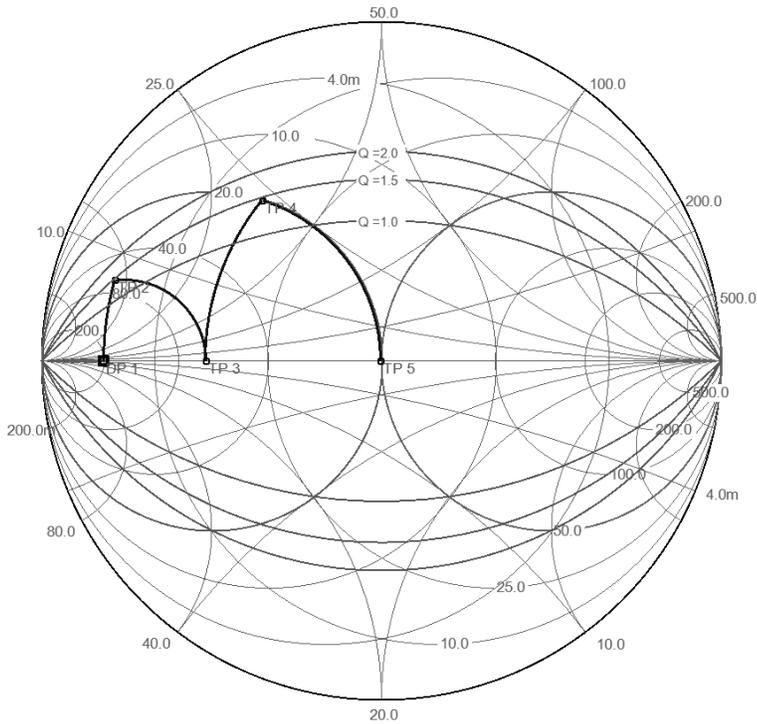


Figure 9 LsCpLsCp LP match

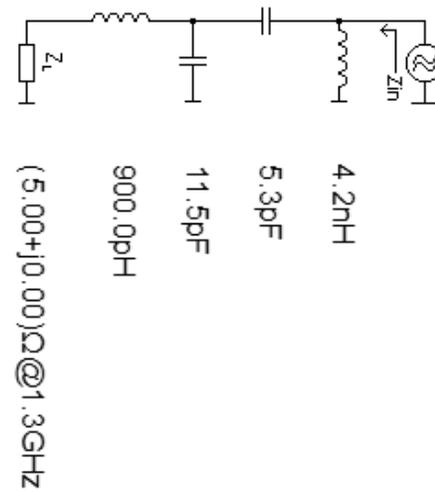
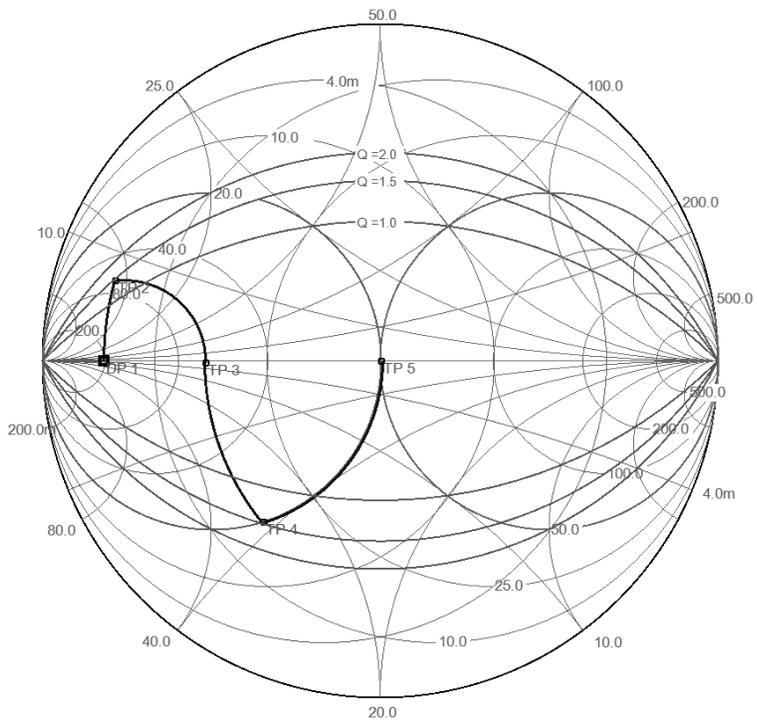


Figure 10 LsCpCsLp BP match

Figures 9 & 10 were generated using the software tool **Smith**:
<http://www.fritz.dellsperger.net/smith.html>.

In general, it is useful to view larger LC ladder structures as cascades of two element LC sections. The loaded Q of each section determines the real transformation (as in a transformer) of that section, by a factor of $[1 + Q_L^2]$. If we wish to limit the loaded Q in the circuit for more bandwidth and lower insertion loss, then more than one section may be required to achieve the overall transformation needed. This is seen in both Figs. 9 and 10 where the two sections allowed the loaded Q to be limited to 1.5. Alternatively, a single section match from 5Ω to 50Ω would require a loaded Q of 3. ($10 = 1 + 3^2$).

Naturally, the question arises as to how many matching elements will yield the lowest loss. That trade-off can be seen in Fig. 11 for a range of transformation ratios and inductor Q. For example, plotted in blue are 2, 4 and 6 element cases with an inductor unloaded Q of 10. The 2 element (single section) curve crosses the 4 element (2 section) case at a transformation ratio of 6. This tells us that it is preferable to use two sections (4 elements) when inductor Q is 10 and the desired transformation is greater than $6x$.

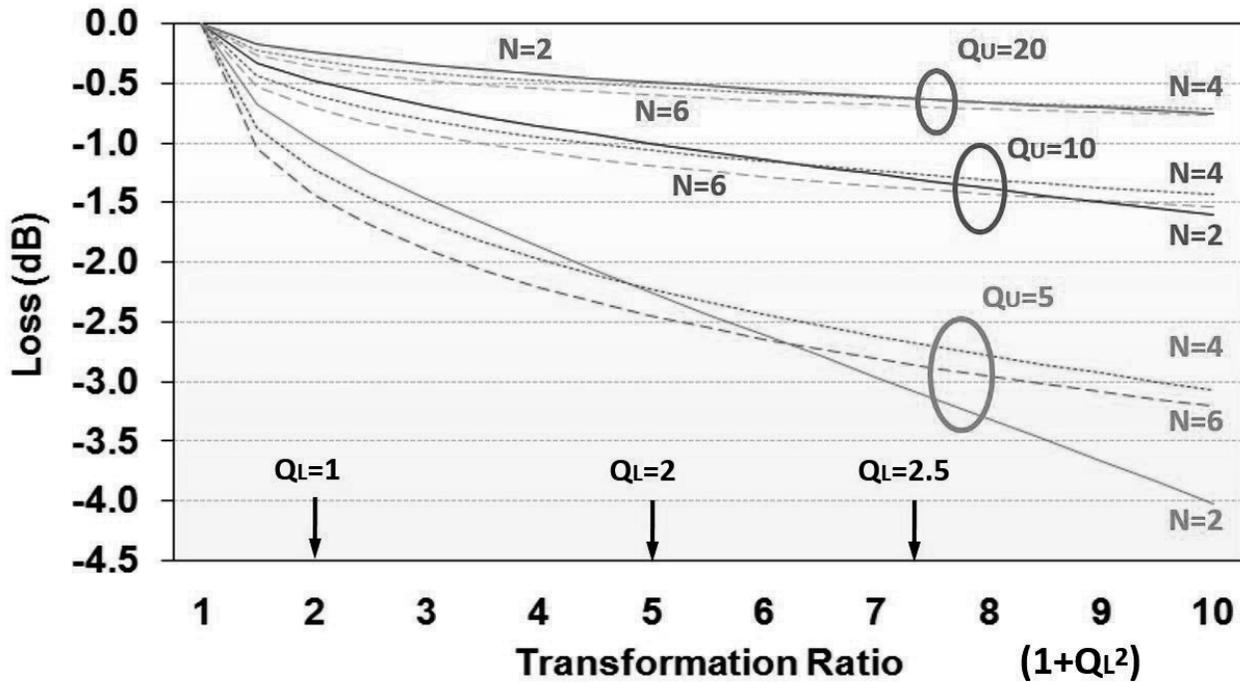


Figure 11 Insertion loss vs transformation ratio and inductor Q

REALIZATION OF A LUMPED DESIGN ON A CIRCUIT BOARD

Let's begin with a bold statement: "There are no truly lumped circuit elements." Chip capacitors have series inductance associated with the package length. Even a small parasitic inductance will eventually resonate the capacitor if the frequency is raised high enough. Similarly, chip inductors have parasitic shunt capacitance that cause a natural self-resonance at some (high) frequency. Chip resistors have series inductance and shunt capacitance due to the non-zero dimensions. Copper traces on a PC board over a back side ground-plane form a quasi-TEM transmission-line called microstrip. That said, its behavior could be approximated by a cascade

of small series inductors and shunt capacitors. The ratio of incremental inductance to incremental capacitance sets the characteristic impedance as: $Z_0 = \sqrt{\frac{L}{C}}$.

The approach here will be to use short ($< 30^\circ$ or $< \lambda/12$) segments of microstrip line to represent series inductors. This is done by using high characteristic impedance where the ratio of incremental series inductance to shunt capacitance is high. Physically, these lines are narrow compared to board thickness.

Short segments of wide traces where the Z_0 is low, represent predominantly shunt capacitance. Pads for chip capacitors fall into this category; so, a circuit node where a shunt capacitor is desired must take into account both the pad capacitance and the effective capacitance of the chip cap. Since, in simplest terms the chip cap is represented as a capacitor in series with a small inductance (due to the body length), the effective capacitance is based on the net X_C of the chip at the operating frequency.

The approach used here is to approximate PC board traces as 3 element LC equivalent circuits. Low Z_0 (wide) lines are represented as a T circuit and high Z_0 (narrow) lines are represented as π circuits. This is a very powerful technique both for estimating chip component parasitics and for realizing printed inductors. This 3rd order lumped LC representation of a length of transmission line is shown in Fig. 12.

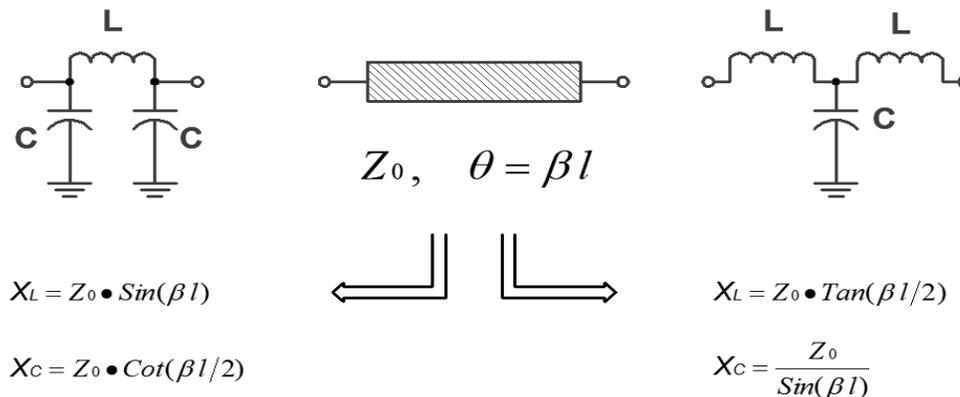


Figure 12 LC representation of a short transmission line

The author has written a software tool that combines the microstrip calculations and the 3rd order equivalent circuit calculations. A windows executable can be downloaded at: <http://k5tra.net/TechFiles/LumpEquiv.exe> . This will make performing the necessary calculations painless. Depending on your windows OS version, you may also need to install vbrun32.dll, a runtime library (usually not necessary on newer PCs). Another useful calculator allows calculation of the effective capacitance of series LC branches (like a chip cap). That tool gives effective capacitance at fundamental, second, and third harmonic frequencies. If a resonance is passed, it will give the effective inductance. This windows executable can be downloaded at: <http://k5tra.net/TechFiles/SeriesTrap.exe> . Figure 13 shows a screen display from both of these tools.

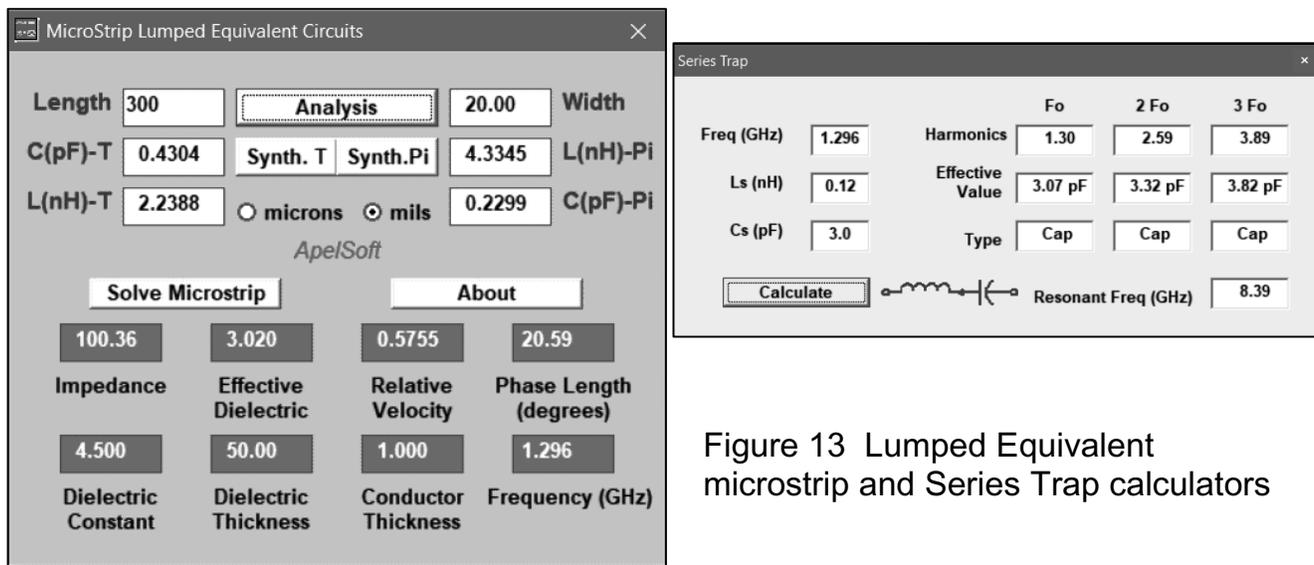


Figure 13 Lumped Equivalent microstrip and Series Trap calculators

Before we work through a design example, it is worth pointing out that a useful estimate of the self-inductance of various surface-mount chips can be made with the microstrip lumped equivalent calculator by loading the package footprint length and width. These results, over a 50 mil FR-4 substrate are shown in Fig. 14. This suggests that an 0603 3pF capacitor is used at 1296 MHz, it will actually look like a 3pF cap in series with 0.77 nH. The Series Trap calculator indicates that the effective capacitance at 1296 MHz is 3.54 pF and at the second harmonic (2596 MHz) it is 7.75 pF. It also indicates series resonance at 3.31 GHz.

CHIP PKG	DIM (mil ²)	L(nH)
0402	40x20	0.59
0603	60x30	0.77
0805	80x50	0.84
ATC 100A	55x55	0.55
ATC 100B	110x110	0.77

Figure 14 SMD series inductance

A DESIGN EXAMPLE

To illustrate the design procedure, let's use a SGA-9189 SiGe transistor to make a 1296 MHz driver stage. This will produce 25 dBm, sufficient to drive a power module. The optimum source and load impedances to present to the transistor base and collector are: $Z_S = 7.1 - j 4.4 \Omega$ and $Z_L = 18.4 + j 4.1 \Omega$. If the matching networks are to provide these optimum impedances, then the complex conjugate impedances can be used as a starting point load in the design of the matching circuits. A reflection match will be a conjugate match. So, the input matching circuit should be designed to interface to a load of: $Z_S^* = 7.1 + j 4.4 \Omega$, and the output matching circuit should be designed to interface to a load of: $Z_L^* = 18.4 - j 4.1 \Omega$.

As a first step, simple lumped LC matching circuits are designed with the complex conjugated optimum impedances as terminations. These can be seen in Figs. 15 and 16.

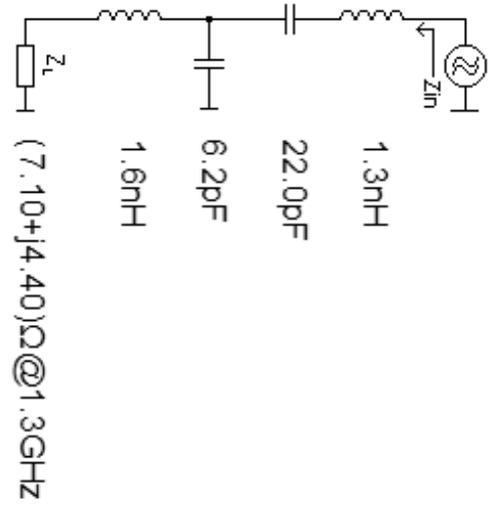
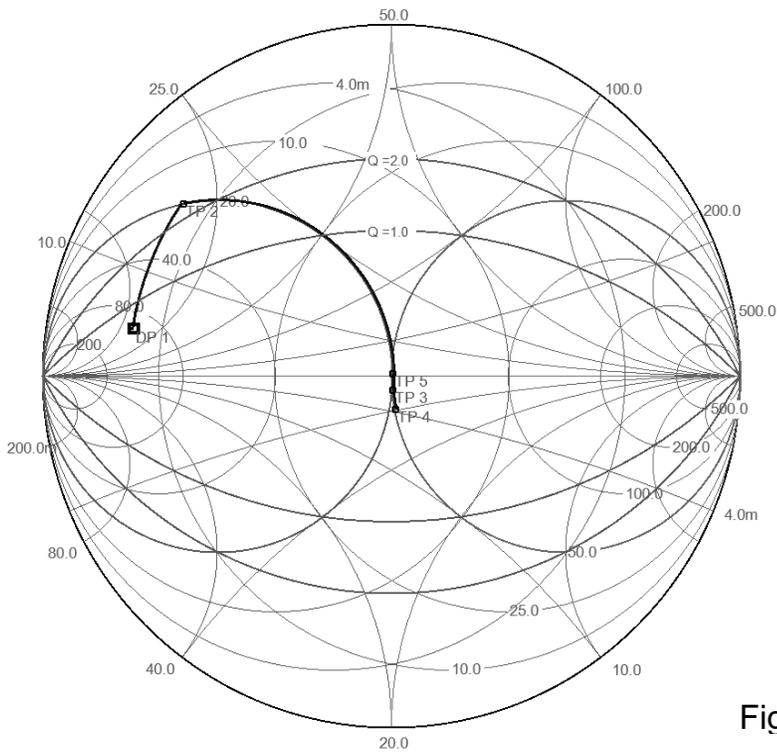


Figure 15 SGA-9189 LC input match

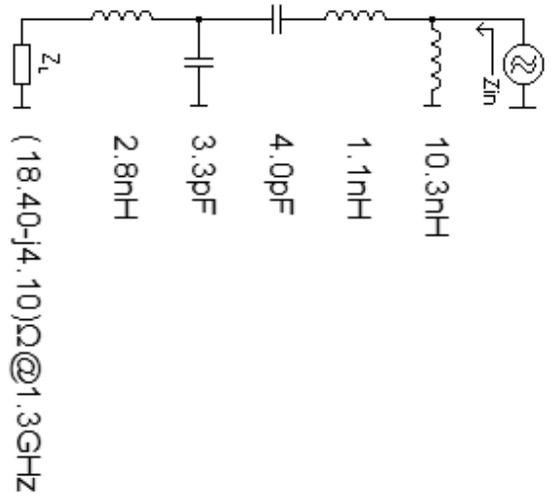
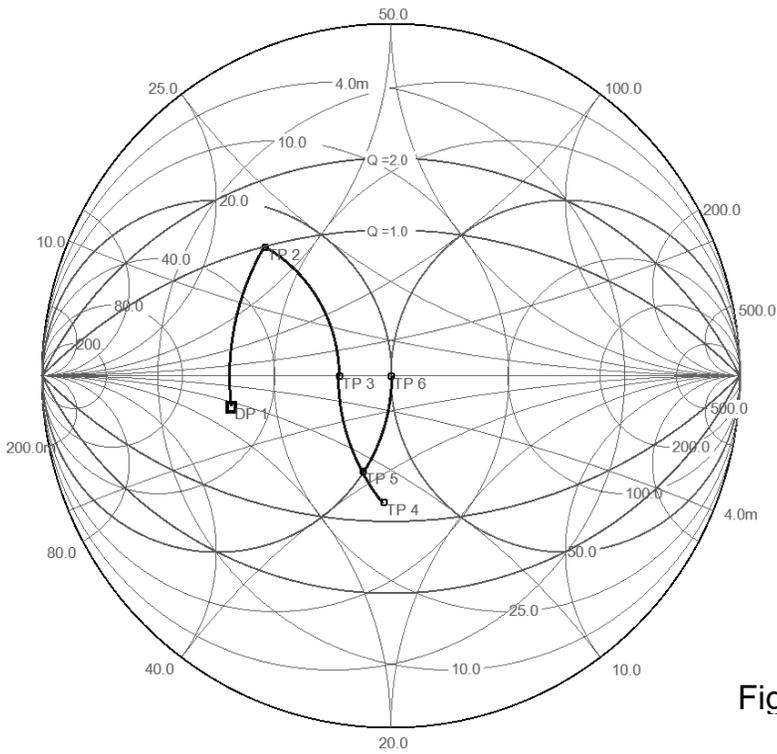


Figure 16 SGA-9189 LC output match

INPUT MATCH ON FR4 BOARD

The input match is a single section L_s , C_p , followed by a 22 pF series DC block capacitor. An additional series inductance around 1 nH was included in anticipation of the parasitic series inductance of the series chip capacitor. The initial series inductance of 1.6 nH can be represented by a 40 mil wide trace on a 50 mil thick FR4 circuit board. This width forms a 77 Ω microstrip line. From the Lumped Equivalent calculator, find that a length of 139 mils is equivalent to a 1.6 nH inductor with 0.136 pF parasitic shunt capacitance at each end. This is shown in Fig. 17.

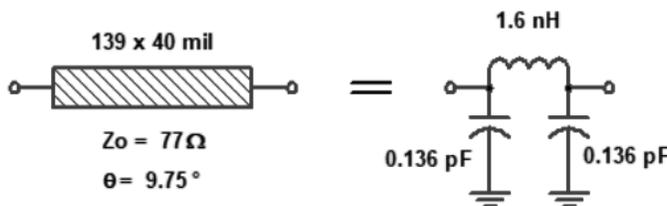
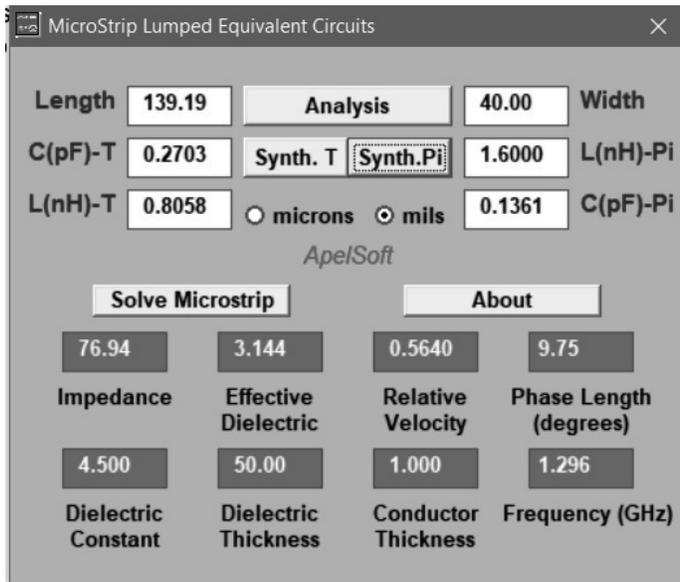


Figure 17 1.6 nH from 77 Ω microstrip

a good solution with a standard value cap and the series inductance that must be absorbed.

To summarize, the input match is comprised of a 139 mil long x 40 mil wide trace, a shunt 0805 3.9 pF chip cap to ground, and a series 0805 22 pF capacitor to the RF input signal (50

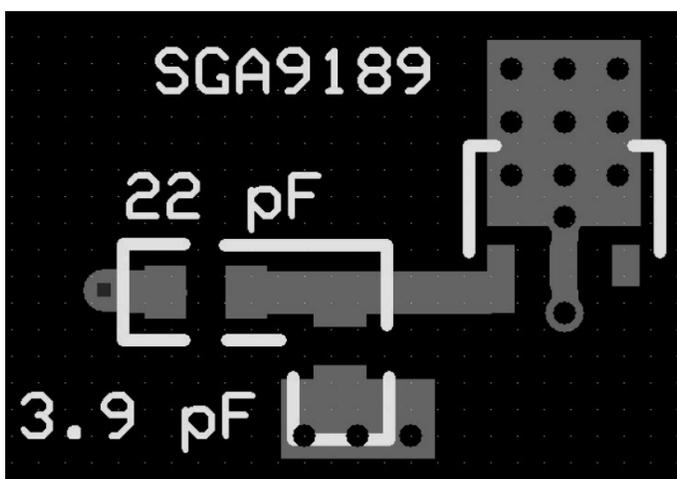


Figure 18 Input RF impedance match

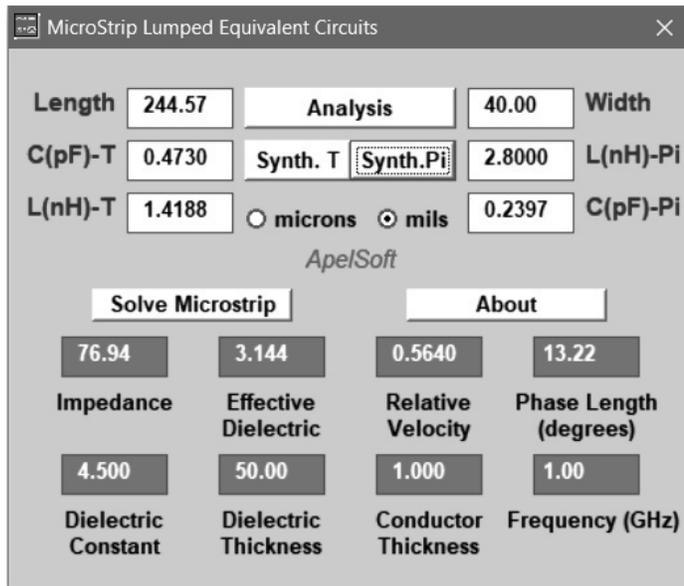
Following the 1.6 nH inductor is a 6.2 pF capacitor (Fig.15). The 0.136 pF parasitic capacitance from the 77 Ω line will account for a very small part of this. Additionally, a shunt chip cap will need to supply the remaining 6.064 pF with its effective capacitance (series LC equivalent). A 0805 chip will have approximately 0.84 nH of self inductance. A 0.035 diameter plate through hole will contribute another 0.52 nH. After a couple calculations with the Series Trap calculator, we find that a 3.9 pF cap in series with 1.4 nH has an effective capacitance of 6.1 pF at 1296 MHz. This is

Ω port). This partial layout can be seen in Fig. 18. Note that the shunt 3.9 pF cap is located in the correct design position, with some position tunability provided in the layout. The region on the transmission line and on the adjacent ground metal where the capacitor could be placed should be defined as exposed 'pad' metal (cleared of solder mask).

OUTPUT MATCH ON FR4 BOARD

The output match is a two section L_s, C_p, C_s, L_p match. The C_s branch provides DC blocking of collector bias voltage. As in the input match, a series inductor around 1 nH is placed in series with the series 4.0 pF capacitor, in anticipation of the self-inductance of a chip capacitor.

The initial series inductance of 2.8 nH, will again be represented by a 40 mil wide trace, 77 Ω microstrip line. From the Lumped Equivalent calculator, find that a length of 245 mils is



equivalent to a 2.8 nH inductor with 0.24 pF parasitic shunt capacitance at each end.

This representation is shown in Fig. 19.

The 3.3 pF shunt capacitor is contributed to by the 0.24 pF parasitic shunt C at the end of the 77 Ω line. So the desired capacitance becomes 3.06 pF. As in the input match, the self-inductance of the shunt chip cap and the ground via hole accounts for approximately 1.4 nH. Once again, the Series Trap calculator yields a 3.09 pF effective capacitance from a series 2.4 pF and 1.4 nH at 1296 MHz.

The next element in the output match is 4 pF capacitance in series with 1.1 nH. After the 0.84 nH self-inductance of an 0805 chip cap is absorbed, a residual 0.26 nH remains to be realized in the layout.

All that remains to be realized is the 10.3 nH shunt inductor. This is considerably more than the previously used inductors. An

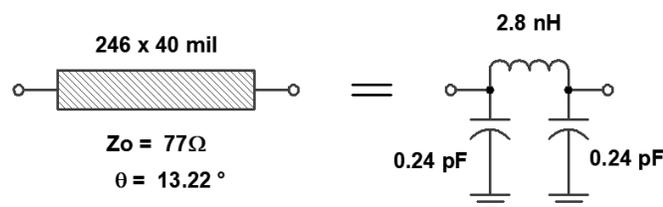


Figure 19 2.8 nH from 77 Ω microstrip

increase in characteristic impedance ($>77\Omega$) of the microstrip line will provide more inductance per unit length. A 10 mil wide line forms a 123 Ω line. The Lumped Equivalent calculator yields a solution of 634 mils long for an equivalent circuit of 10.3 nH with parasitic shunt capacitances of 0.39 pF at each end. That much shunt capacitance across the inductance will produce an effective inductance that is around 14 nH; so we will need to reduce the length of this line. This effective inductance of an LC parallel connection is analogous to the effective capacitance we encountered earlier in series LC branches representing a capacitor. That said, you will probably also want to download the ShuntTrap calculator at:

<http://k5tra.net/TechFiles/ShuntTrap.exe>. This tool will allow you to easily find that 8.67 nH in parallel with 0.28 pF is equivalent to 10.3 nH at 1296 MHz. The Lumped Equivalent calculator provides the physical solution for this as: 470 mils x 10 mils (8.15 nH). The total 8.67 nH = 8.15 nH + 0.52nH via inductance. This solution is shown in Fig. 20.

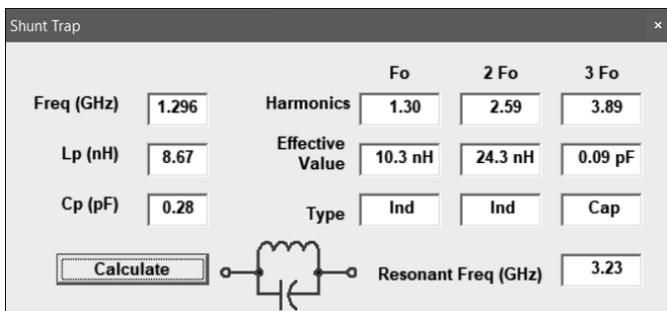
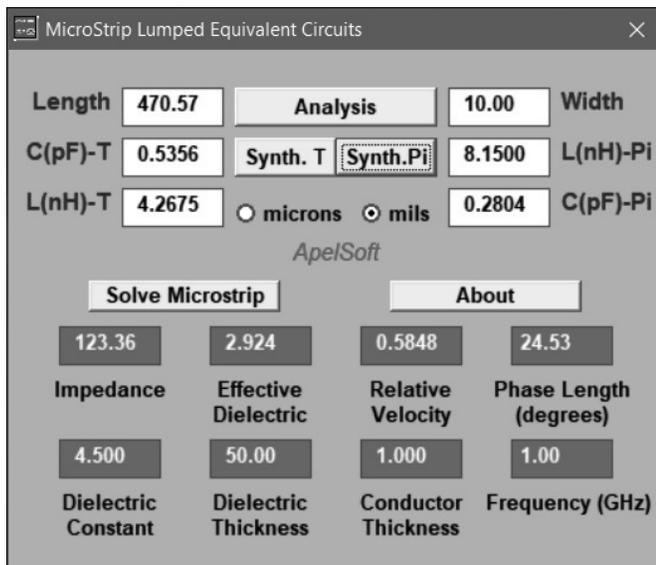
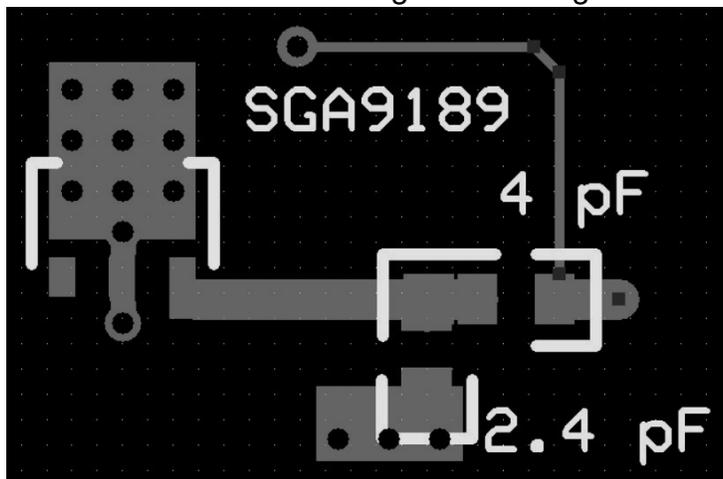


Figure 20 10.2 nH solution with Lumped Equivalent and Shunt Trap calculators

In summary, the output match is comprised of a 246 mil long x 40 mil wide trace, a shunt 0805 2.4 pF chip cap to ground, followed by a series 0805 4 pF capacitor and a shunt 470 mil x 10 mil transmission line to ground through a 35 mil via (total approximates 10.3 nH). The RF



output input signal (50 Ω port) is at the junction of the 4 pF series cap and the shunt 478 mil line. This partial layout can be seen in Fig. 21. Note that the shunt 2.4 pF cap is located in the correct design position, with some position tunability provided in the layout. The region on the transmission line and on the adjacent ground metal where the capacitor could be placed should be defined as exposed 'pad' metal (cleared of solder mask).

Figure 21 Output RF impedance match

CONCLUSIONS

A methodology for realizing board level impedance matching circuits has been presented including a design example at 1296 MHz. This methodology includes utilizing board microstrip traces to approximate inductors from a lumped LC design. Additionally, a methodology to design lumped LC matching circuits with the Smith Chart was presented. An overview of the Smith Chart was also presented as an initial part of that discussion. Load Q and the $Q \cdot BW$ product was briefly discussed, due to the importance of Q_L in setting the limits of achievable impedance match performance. A number of software tools were highlighted in this paper. The K5TRA software tools are free and downloadable from <http://k5tra.net>. The Smith tool is not

free; but it is well worth the small expense. Smith is available from:
<http://www.fritz.dellsperger.net/smith.html>.

The overall layout of the SGA9189 amplifier is illustrated in Fig.22, including bias feed lines. Some degree of tunability was built-in to the layout. Line junction and step discontinuities were not included in this paper, as they were minimal in the example design and outside the intended scope of the paper.

It is hoped that the approach presented here will stimulate more board level RF home brewing.

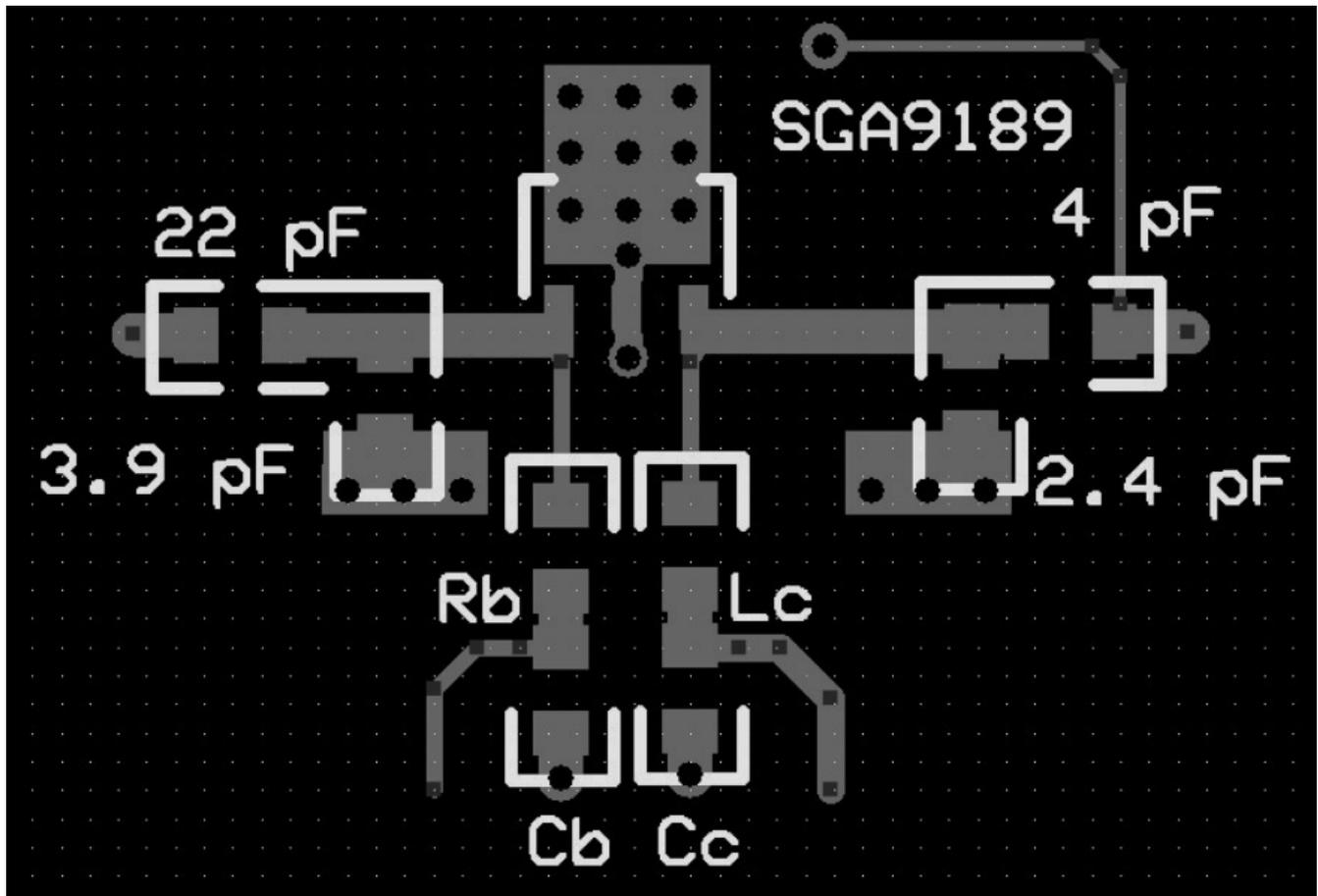


Figure 22 Overall SGA-9189 amplifier layout

10 and 24 GHz Terrestrial DX

by Rex Moncur, VK7MO

The use of portable stations designed for EME (20 to 50 watts feeding 70 cm to 1 metre dishes) using digital modes and GPSDO locking opens up opportunities to increase terrestrial distances at 10 and 24 GHz, explore various modes of propagation and the effects of spreading and absorption. This has led to World Record distances of 2732 and 2793 km for terrestrial propagation on 10 GHz in southern Australia⁽¹⁾⁽²⁾.

Techniques for setting accurate bearings such as differential GPS ⁽³⁾ that have been refined for EME are also of considerable value for terrestrial operations. Single tone techniques developed for EME are useful for looking for terrestrial openings. Atmospheric absorption is a critical factor for not only 24 GHz but also for long distance terrestrial contacts at 10 GHz. For example, analysis of the 10 GHz 2732 km contact shows absorption losses of around 40 dB and explains why 10 GHz signals were so much weaker than UHF at the same time.

Modes of Propagation and Reported DX

Propagation	10 GHz	24 GHz
Tropo-scatter*	650 km Estimated*	250 km Estimated*
Aircraft-scatter	905 km VK7MO to VK3HZ	566 km VK7MO to VK3HZ
Rain-scatter	1129 km DK3SE to IK7UXW	710 km LX1DB to F2CT
Tropo-ducting	2793 km VK7MO to VK6DZ	581 km DL7QY to F6DKW

Fig 1: DX Records as reported on the OK2KKW web site. (* Tropo-scatter is not generally reported separately from tropo-ducting so estimates are included based on tests between VK7MO and VK3HZ)

Meteorological Data for understanding DX situations

Excel Spreadsheets have been developed by the author to show the intensity and height of ducts and also the absorption loss based on radiosonde data. Radiosonde data is widely available for sites throughout the World and is made available by the University of Wyoming as below:

University of Wyoming Upper Air Soundings web site:

<http://weather.uwyo.edu/upperair/sounding.html>

Spreadsheet 1: Ducting

Spreadsheet 2: Atmospheric absorption

Please use the link below to download the spreadsheets from a Google Drive directory; then open to use on your computer. The method of opening may depend on your operating system and browser. On Windows with Firefox you right click on the spreadsheet picture and then go to download. The graphs work correctly on Microsoft Excel but may not on other programs. For example on Google Sheets the graphs are changed from lines to dots and are not useful for finding ducts.

https://drive.google.com/open?id=1ctR-ptUrW_6ZBmOISkxy2NpUrM1Su_Gz

If you wish to try out these spreadsheets an interesting example is to try the Hawaii to California path. Set the date to 29 July 1991 at 00z and click on PHTO at the Hawaii end to find the data. Copy this data into Spreadsheet 1 and you will see an intense duct at around 2500 metres that goes off the scale. This relates to the time of the still current 5.7 GHz terrestrial World Record between N6CA and KH6HME. You can use data at the California end at San Diego (NKX) to see an interesting change in the level of the duct. You can use Spreadsheet 2 at the height of these ducts to gain estimates of the absorption that these stations had to cope with at 5.7 GHz and how much worse it would have been at 10 GHz.

Analysis of 2372 km 10 GHz QSO based on the above Spreadsheets

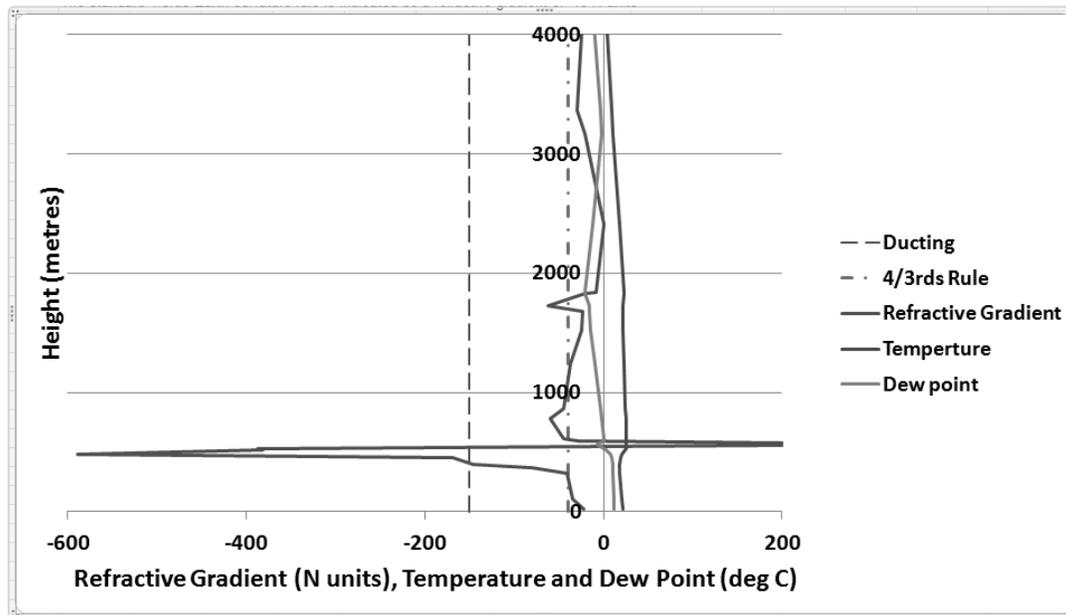


Fig 2: Shows radio refractive gradient at the time of the 2732 km 10 GHz terrestrial QSO on 5 January 2015.

Fig 2 shows an intense duct at about 500 metres with a refractive gradient of -588 N units. Note that ducting can promulgate around the curvature of the Earth with a refractive gradient of just -157 N units indicating that this was indeed a very intense duct. While absorption losses are much higher at 24 GHz they cannot be ignored on long terrestrial paths at 10 GHz. Applying Spreadsheet 2 at the time of this QSO gave 40 dB absorption loss at 10 GHz at 500 metres and at least in part explains why 432 MHz was S9 compared to 10 GHz at S1 (if one takes account of relative propagation loss, antenna gains, and power, 10 GHz should have been in front by around 10 dB).

Fig 3 is an International Telecommunications map of the global incidence of ducting. Surprisingly the current world record across Southern Australia is not in an area of high incidence nor is the Hawaii to

California path. There are a few possible explanations. Firstly, records are more driven by availability of operators and locations, and secondly, while ducting tends to be more frequent in the tropics, absorption which affects 10 GHz is much higher in the tropics. There are some interesting possibilities from the East Coast of Australia out to the Central Pacific but absorption is likely to be a limitation.

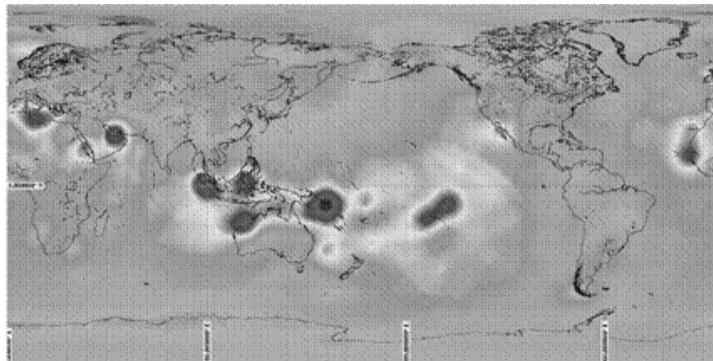


Fig 3: Global incidence of ducting

Atmospheric Absorption

Atmospheric absorption is the key reason that 24 GHz is so much more difficult than 10 GHz (Fig 4).

Temperature and Relative Humidity	1.296 GHz (dB)	10.368 GHz (dB)	24.048 GHz (dB)
0 Deg C, 20%	3.4	5.1	18.9
0 Deg C, 50%	3.4	5.9	34.9
0 Deg C, 80%	3.5	6.6	50.8
15 Deg C, 20%	3.0	5.2	35.0
15 Deg C, 50%	3.0	7.1	76.4
15 Deg C, 80%	3.1	9.0	117.1
30 Deg C, 20%	2.7	6.2	71.0
30 Deg C, 50%	2.7	10.4	164.0
30 Deg C, 80%	2.8	14.8	254.6

Fig 4: Absorption at sea level, over a 500 km path based on International Telecommunications method which has been applied with a Spreadsheet 2.

Atmospheric absorption at 24 GHz is primarily related to water vapor, which is in turn directly related to "Absolute Humidity". A common misunderstanding is to use "Relative Humidity" as an indicator of absorption rather than "Absolute Humidity". As shown in Fig 3 the simple use of "Relative Humidity (RH)" without referring to temperature can lead one to think that it is better to operate at 20% RH and 30 degree C whereas in fact the loss would be 20 dB less by operating at 80% RH and 0 degrees C .

Unfortunately it is not easy find a forecast of "Absolute Humidity" to plan a DX attempt. However, forecasts of Precipitable Water are readily available up to 4 days in advance. Most of the Water Vapor is concentrated in the lower layers of the atmosphere and while Precipitable Water is the sum of all

layers it is a good relative approximation of what one can expect and relates directly to the absorption caused by water vapour. URLs to find forecasts of Precipitable Water, are as follows:

North America: <http://wxmaps.org/pix/nam.pw>

Australia: <http://wxmaps.org/pix/aus.pw>

Europe: <http://wxmaps.org/pix/euro.pw>

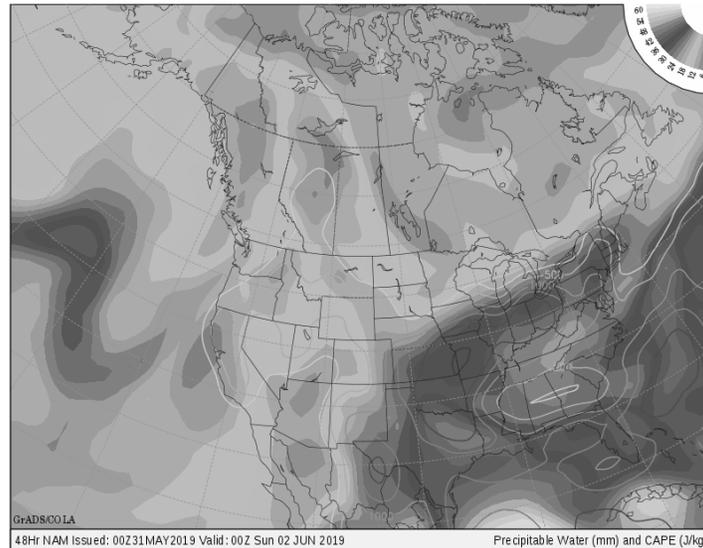


Fig 5: An example of a Precipitable Water forecast for North America.

From the example at Fig 5 it is seen that the water vapor can vary from around 5 mm in Canada to 50 mm in Dallas which means the absorption loss due to water vapor is around 10 times worse in Dallas at the time of this forecast.

Absorption Loss on Aircraft Scatter

One means of reducing absorption is to use aircraft scatter so that much of the path is at high levels where the water vapor is low. This essentially eliminates absorption at 10 GHz. On 24 GHz we have been able to complete an aircraft scatter QSO out to 566 km but despite repeated attempts at 700 km we received nothing at all while 10 GHz was giving strong signals. We have concluded that at the low elevations necessary to work to 700 km so much of the path goes through the lower atmosphere that the absorption losses prevent a QSO.

Absorption Loss on Rain Scatter

Rain-scatter can occur at potentially higher levels than the flight levels of civil aircraft and thus reduce the absorption losses and potentially allow longer distances. In addition the scattering volume is typically much larger which should at least partly offset absorption losses. Our experience on 10 GHz is that on a 580 km path rain-scatter can be significantly stronger than aircraft-scatter. A potential issue with rain-scatter at 10 and 24 GHz is the loss of signal due to scattering from rain at low levels

along the path but we have not seen this as significant on 10 GHz. My preliminary view is that loss due to rain along the path is not all that significant as the intercepting rain drops still scatter useful energy in the forward direction. This is a fertile field for further work on 24 GHz with EME capable stations.

Spreading

Spreading affects rain-scatter and tropo-scatter as in both cases the scattering is from a large common volume and the path length varies from the side of the beam to the centre. Spreading can be significantly reduced by using forward scatter rather than seeking out strong rain cells to the side. Forward rain-scatter works well with low spreading if both stations beam accurately at each other and just wait for the rain to pass between. Even light rain that is not visible on a 4 GHz rain radar is sufficient to complete a digital QSO on 10 GHz over 600 km. QRA64 D works well for both rain and tropo-scatter at 10 GHz with dishes in the range 60 cm to 1 metre.

Forward scatter spreading increases in proportion to antenna beam-width and also in proportion to frequency. However, if one uses the same size dish then the beam-width reduces in proportion to frequency and compensates for the increase in spreading with frequency. Thus for the same size dish forward scatter spreading is similar for 10 and 24 GHz and is typically in the range of 10 or 20 Hz and QRA64D or E should work well.

Aircraft scatter is essentially from a point source and does not suffer from spreading.

Interestingly, our experience shows that tropo-ducting also does not suffer from spreading which suggests that the path length does not vary. Prior to the onset of ducting we have seen spreading of several Hz, which we put down to a combined tropo-scatter/tropo-ducting path. But once the path is fully tropo-ducting the spreading is less than 1 Hz and even the QRA64A sub-mode can be used.

Operational and Planning Issues

The secret to finding a weak opening such as a tropo-duct or aircraft scatter is to remove as many variables as possible while you wait for hours or more for the signal to rise out of the noise. Both stations must be exactly on frequency with GPSDO locking and both must be accurately aligned in azimuth and elevation. In looking for a duct to open there is advantage in using the integrated single tone facility on WSJT-X in conjunction with QRA64 - single tones are transmitted as 1000 Hz by ticking the "sh" box and using TX box 6.

You look for the integrated signal tones on the Spectrum Graph, set to linear average which gives around 6 dB more sensitivity than exchanging messages. To be sure you are seeing a real signal it is useful to have both stations within a few Hz which means locking both the 10 GHz transverter and the IF transceiver with good quality double oven GPSDOs. To ensure that both stations are accurately aligned we use a rifle scope which has been accurately aligned to the dish on sun noise.

In the past we have aligned on some geographic feature as a reference and used say Google Earth to find its bearing and then moved to the bearing of the other station with an accurate 360 degree

protractor which is calibrated on the reference . However, if beaming over the sea it is sometimes not possible to find a suitable reference and our recent development of differential GPS allows accurate alignment to better than 0.1 degrees on just a 5 metre baseline and it thus much preferred ⁽³⁾.

It is important to be sure both systems are reliable and working before you spend hours waiting for an opening. In this respect it is useful to check cold sky to ground or sun as a check on receive system performance. To ensure both stations are in fact GPSDO locked it is useful to have an independent GPSDO locked weak signal source. To ensure one is in fact transmitting, one can monitor the PA current.

Hepburn Charts are important to plan portable operations when ducting is more likely to be present. For rain scatter, access to meteorological radar data on the web is important. For aircraft scatter we take an ADSB receiving system to monitor aircraft positions. We normally beam directly at each other and wait for the aircraft to pass between - but occasionally to get a rare grid locator it might be necessary to use side scatter (which is weaker) and both stations beam directly at the aircraft.

To date our 10 GHz tropo-ducting and aircraft scatter records have all been done with a 60 cm to 77 cm dish and 10 watts at one end⁽⁴⁾. On 24 GHz we have used only 4 watts at one end to a 60 cm dish and a 113 cm dish at the other. My more recent EME system runs 90 watts to a 113 cm dish on 10 GHz and 20 watts to a 113 cm dish on 24 GHz and if similar capability was available at both ends this should provide a useful improvement. All PAs are solid state for safe operation in the field.

Improvements in System Performance

Back in 1994 VK6KZ and VK5NY set the then 10 GHz terrestrial World Record across southern Australia at 1912 km. They used 40 cm dishes, 100 mW and SSB with drifting oscillators and used VHF signals align their antennas. With today's EME capable portable stations we can use 60 watts (+28 dB), 1 metre dishes at both ends (+16 dB), and digital modes (+ 30 dB) -- an improvement in system performance of over 70 dB.

Acknowledgements

To VK3HZ who has not only been at the other end of many contacts but also contributed to most of the development that has led to the equipment and techniques which are the subject of this paper. To VK6DZ who was at the other end of the world record terrestrial QSO's at 10 GHz. To VK7WLH for reviewing this paper.

CONCLUSIONS

EME capable digital and GPSDO locked portable stations offer good prospects for extending 10 and 24 GHz records with modes such as QRA64D. The single tone integration feature on WSJT-X can add around 6 dB sensitivity in finding the presence of a weak signal while waiting for an opening.

Ducting is the best prospect for 10 GHz but note that as the distance increases absorption losses become significant - particularly in tropical areas and it may be better to operate in temperate areas even if the openings are less frequent.

Forward rain scatter appears to be the best prospect at 24 GHz and digital modes should significantly enhance the capability.

References:

- (1) Rex Moncur, VK7MO, et al, "10 GHz across the Great Australian Bight & New World Record (5 Jan 2015) DUBUS 2015 Volume 1 page 117.
- (2) Rex Moncur, VK7MO, et al, "Extension of the 10 GHz terrestrial World Record to 2793 km", DUBUS 2016 Volume 1 Page 115.
- (3) David Smith, VK3HZ & Rex Moncur VK7MO, " Differential GPS Azimuth Reference for Microwave Portable Operations" . Paper included with MUD proceedings 2019.
- (4) Rex Moncur, VK7MO, and David Smith, VK3HZ, "Aircraft Scatter on 10 and 24 GHz using ISCAT", DUBUS 2014, Volume 4, Page 92.

78 GHz DX'ing in the Desert Southwest

By Barry Malowanchuk, VE4MA

As a winter resident of Arizona for a number of years now, I have heard many stories of the great long distance 10 GHz QSOs from Arizona Mountain tops. The only regular activity on 10 GHz has occurred during the ARRL "10 GHz & Up" Contest in the late summer, when the temperatures are very high and with high humidity that severely affects the bands above 10 GHz. During the winter months the Arizona temperatures are much lower and the atmosphere very dry and thus more supportive of long distance QSOs at 24, 47 and 75 GHz. This paper will discuss efforts to explore the possibilities for real DX on 75 GHz.

Distance Records

The records for microwave frequencies are kept by the ARRL at www.arrl.org and for early 2019 the records for 47 and 75 GHz are shown in Figure 1.

47 GHz			
Tropo (C)	344.8	AD6FP /6 (DM07as) - W6QIW/6 (DM04ms)	19-Sep-15
Tropo (C)	343	W6QI/6 (DM04ms) - AD6FP/6 (DM07bs)	30-Oct-05
Tropo (C)	246	W0EOM/6 (CM97ei) - KF6KVG/6 (DM06ms)	9-Feb-03
Tropo (C)	232	KB8VAO/6 (CM96qi) - AD6FP/6 (DM04ms)	18-Sep-04
Tropo (C)	215	KT1J/N1JEZ (FN34bi) - VE2UG/VE3FN (FN26rf)	30-Jun-16
Tropo (C)	205	N1JEZ/1 (FN44ig) - WA1MBA/1 (FN42bl)	3-Jul-05
Tropo (C)	174	W3IY/4 (FN10ff) - W4SW/4 (FM08us)	14-Nov-01
Tropo (C)	120	NU7Z (CN87ms) - KD7TS (CN96aw)	4-Oct-03
75 GHz			
LOS	289	AD6IW (CN90fl) - KF6KVG (CM97av)	23-Jun-14
LOS	289	AD6IW (CN90fl) - K6GZA (CM97av)	23-Jun-14
LOS	205	N1JEZ/1 (FN44ig) - WA1MBA/1 (FN42bl)	19-Sep-14
LOS	177	AD6FP/6 (CM88qp) - KF6KVG/6 (CM97ad)	1-Mar-02
LOS	177	W0EOM/6 (CM88qp) - KF6KVG/6 (CM97ad)	1-Mar-02
LOS	110	K2AD (EM96ur) - W2SZ (FM07fm)	20-May-99

Figure 1 ARRL 47 & 75 GHz Distance Records March 2019

Notably the distances for both bands are similar in spite of the fact that 47 GHz equipment has enjoyed the benefit of possibly having receiver and transmitter amplifiers for a number of years now. Amplified equipment has only become available recently for 75 GHz. The distance limitation is mostly imposed by the requirement for Line of Sight (LOS) visibility between stations. Of course with low powers and poor receivers the option of going beyond the horizon has not been an option.

75 GHz Propagation

On Line of Sight paths the path loss is defined as $20 \log (4\pi d/\lambda)$, which can be calculated. At frequencies above 10 GHz there are atmospheric absorptions caused by Oxygen and Water Vapor in the air. Brian WA1ZMS has generated a convenient set of curves for the microwave bands from 24 to 411 GHz and are available at www.wa1mba.org. The curve for 76 GHz is shown in Figure 2 below.

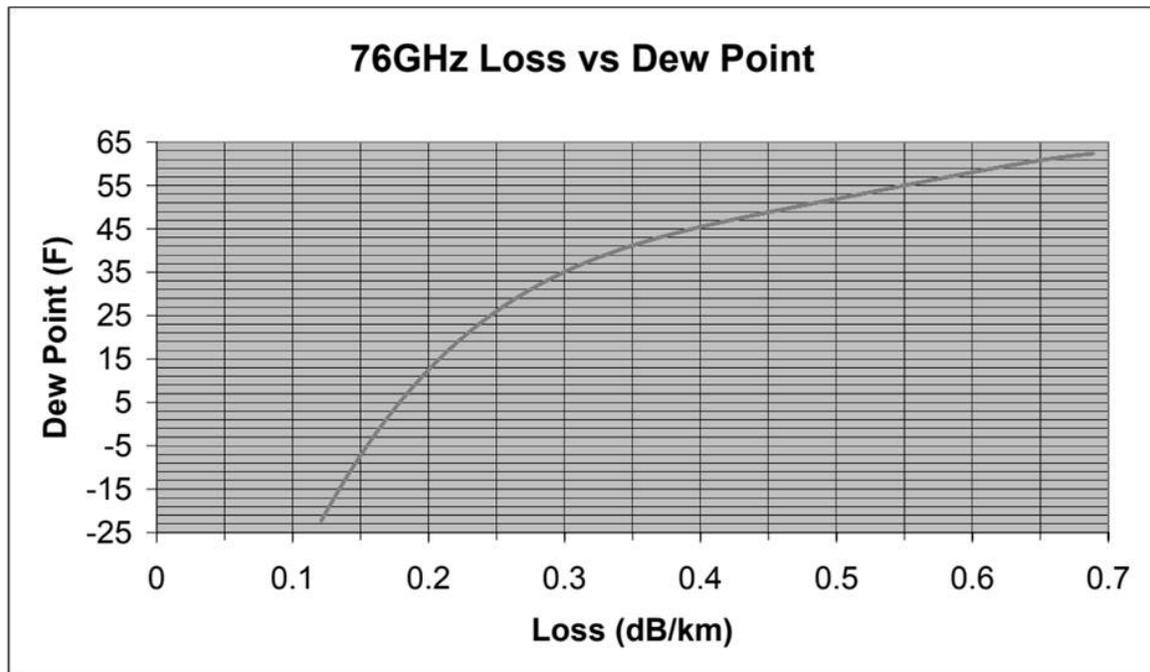


Figure 2 WA1ZMS Atmospheric Loss Chart for 75 GHz

It can be seen that even at very low “dew point” temperatures, the additional path loss due to atmospheric is greater than 0.1 dB / km. Thus on a 200km path that is an additional 20 dB of loss. This loss increases dramatically as the temperature increases above freezing (32 deg. F) to 55 dB! Fortunately as the altitude increases in order to gain a longer LOS distance, the temperature normally drops as well as atmospheric pressure, so this tends to lower the calculated atmospheric losses. So a good Line of Sight and a very low Dew Point temperature over the length of the path are essential for very long distance QSOs.

Path Selection

The two (2) most important aspects of planning 78 GHz path tests are of course “Locations” (the LOS path) and the weather forecast. I was indeed fortunate to have some veteran Arizona 10 GHz operators available to provide advice on locations that could be used for long distance tests. Unfortunately 10 GHz is very much like a VHF band in that it propagates very well over the horizon (non-LOS paths) and stations generally have a much higher ERP than is possible on the mmWave microwave bands.

Once a list of possible paths was developed it was necessary to evaluate each path with software to see if the paths were LOS or obstructed. We know that at 10 GHz obstructed or knife edge paths can work, but we have NO experience for anything like this at 24, 47 or 78 GHz. Several software

packages were used including freeware like “Radio Mobile” and “Hey Whats That” and a commercial product. The freeware products provided conflicting results for marginal LOS paths and so they were evaluated with the commercial software thanks to Brian WA1ZMS. “Hey Whats That” does provide good results however and the learning curve is not so steep on its use. It provides a list of all points visible with the bearing and distance.

Initially we wanted a short hop, just something to check out the equipment on locally and then at a short distance (~60 km). There are mountains that ring around the City of Phoenix, with many hiking trails and good access, so that it was not a problem to find a 60 km hop.

Of course I was looking for the ultimate DX path...one that would break the existing record of 289 km but I was disappointed. There are many hops in the 265 km range and some that are close to 280 km but are subject to the uncertainty of the LOS.

There is a hop in Southern Arizona from Mt. Lemmon (near Tucson) to White Tanks (Just west of Phoenix) and the distance is 207 km (120 mi). This hop is covered on 1.2 and 2.4 GHz for linking digital ATV repeaters, using a 2.4 m (8ft) dish. The path profile is shown in Figure 3.

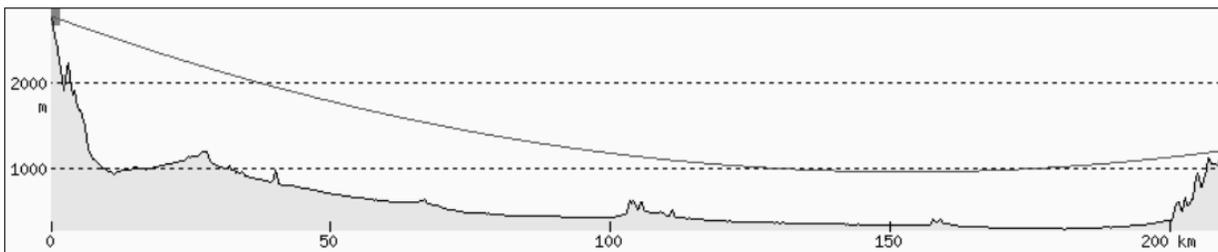


Figure 3 Mt Lemmon to White Tanks 207 km Path Profile

A longer 265 km hop exists from Mt Lemmon to Mt Union (NW of Phoenix near Prescott) and the path profile is shown in Figure 4.

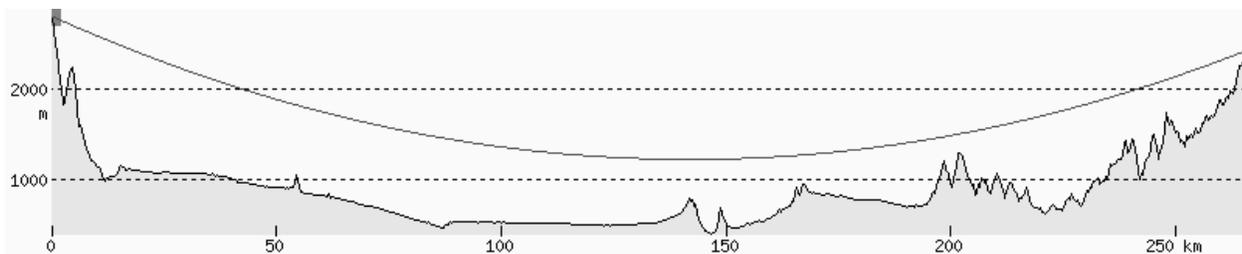


Figure 4 Mt Lemmon to Mt Union 265km Path Profile

As mentioned earlier the ability to knife edge or go over the horizon is unknown as yet. A 280km hop exists from Mt Lemmon to Mt Harquahala which has some obstruction (~ 10 m or 35 ft). This hop was studied closely and the most optimistic path profile used the commercial software and is shown in Figure 5. Other similar slightly obstructed hops exist elsewhere that could be record breakers.

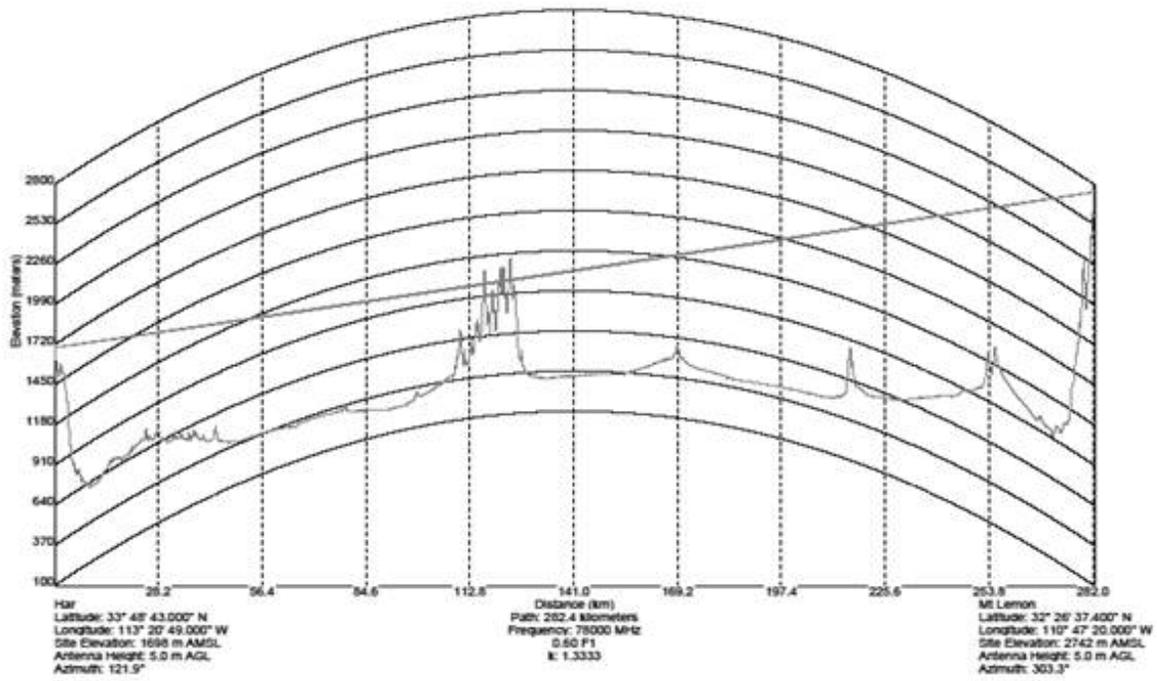


Figure 5 Mt Harquahala to Mt Lemmon 280 km

Arizona Weather

Of course we know “It Never Rains in Arizona”...at least that is the way IT WAS BEFORE 2019. Usually weeks would go by with sunny dry weather. In the winter of 2018/2019 The amount of rain was twice (2X) the normal amount and it seemed like it rained about every third day. The Historical Annual Humidity levels for Phoenix are shown in Figure 6. It shows that the weather should be dry from December until near the end of May.

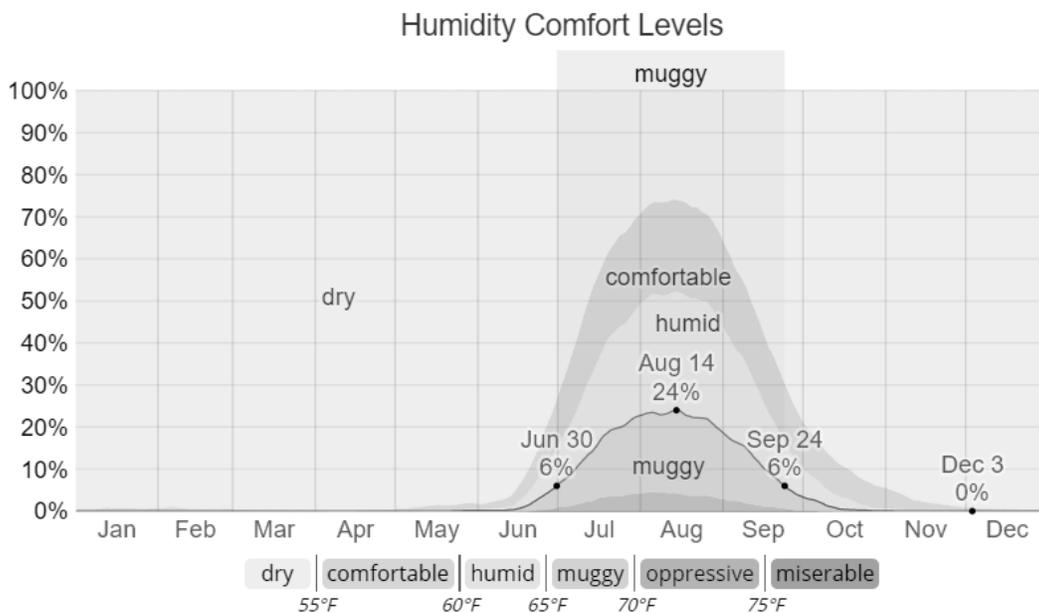


Figure 6 Historical Annual Humidity Levels for Phoenix

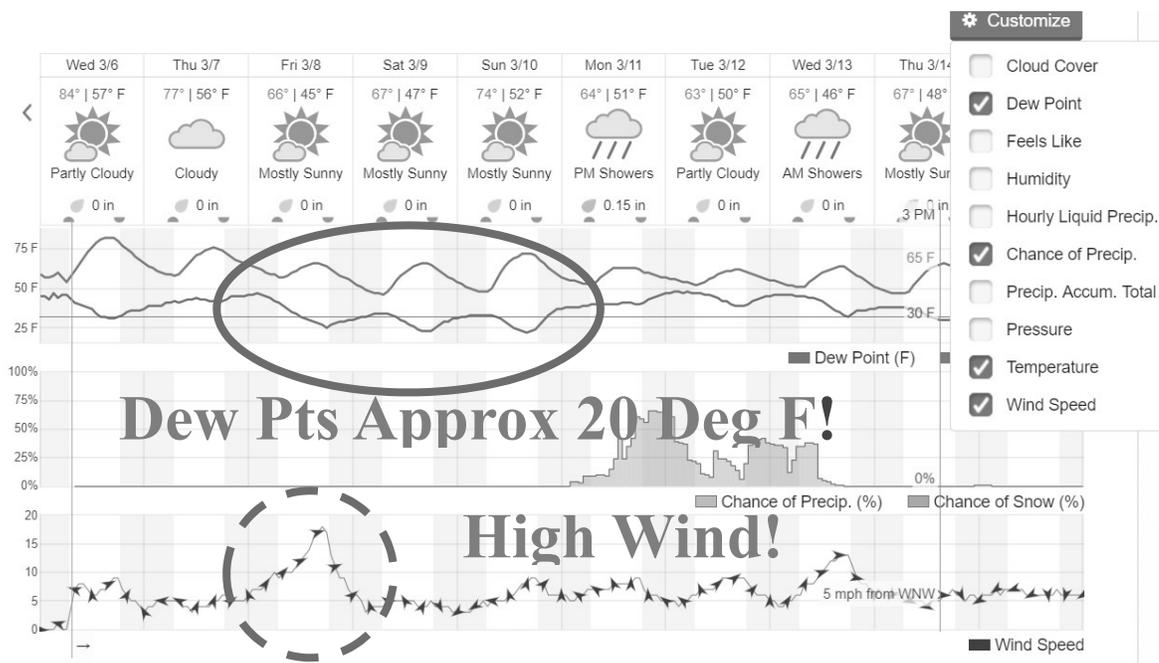


Figure 7 Phoenix Weather Forecast for Early March 2019

The weather was so unsettled in January & February and not looking very good at the beginning of March. The Dew Point temperatures were just not coming down below 30 degrees F. I began looking at the forecasts from “Weather Underground” and they were showing a period of 3 days with very low (20 deg. F) Dew points for Phoenix (see Figure 7). The average Dew Point temperature for a hop has to be “guess-timated” since the information is not available for all locations. Phoenix being the approximate low point in the hop should have the greatest influence, but the end points at high altitude should have values much lower.

It is notable that the Dew Point was shown to be lowest on March 8, 9 & 10 with a return to the rainy weather pattern starting on Monday March 11. It also showed a period of very high winds on Friday March 8. Notably this weather forecast remained unchanged as the days approached. These dates fit in well with the availability of other stations and me. The group was scheduled for these dates.

The Operating Groups & Equipment

With the short notice of this activity not everyone was able to attend. Two groups were assembled, the first one with Mark N0IO from Colorado, Bill W7QQ from New Mexico and who were escorted by Kevin AD7OI & Tammy KI7GVT. The second group included Barry VE4MA, Al W5LUA from Texas, Tony K8ZR from Ohio, Jim K0KFC from Wisconsin and who were escorted by Steve KJ7OG and Ron K7RJ and his wife Clare.

The organization of the groups was so that all bands from 10 to 78 GHz were available at both ends. We also were able to make use of a linked 2m repeater for liaison although 10 GHz would prove valuable in this regard as well as for optimizing antenna pointing.

The “75 GHz” amateur band was expanded in 2017 to include the spectrum from 76 to 81 GHz, which is the spectrum used by vehicular radar system. For this event we chose to operate at 78192 MHz

since 3 of the rigs did not have frequency agility and had been configured for this. The path loss is actually lower at 78 GHz than at 76 GHz where most of the world operates. The equipment for 78 GHz was very important since we were expecting the most difficulty to cross long paths at this frequency. Four of the Five Stations incorporated WA1MBA receive preamplifiers giving an approximate 5 dB noise figure. All stations were capable of high power (0.25 to 1 W) but by the 2017 license restriction we adjusted our TX powers to produce the allowed 55 dBm (315 W) ERP. Three of the stations used 30 cm (12 in) dishes with 10 mW of power while the fourth station uses a lower gain long horn and correspondingly more TX power for the same ERP.

Operating Results Day #1 Friday March 8, 2019

We began the day by reconfiguring K8ZR's station to operate on 78 GHz. Tony's earlier work in Ohio and Michigan was on 76032 MHz. This was easy to do with his programmable Kuhne PLL Lo and Image Rejection Kuhne transverter. We then did a basic check to ensure that all rigs were operating before we moved to our first operating position near the San Tan Mountain. The second group moved to Shaw Butte in North Central Phoenix, which gave us an LOS path of 63 km (39 mi). See the path profile in Figure 8. Signals were very strong with SSB used on all bands through 78 GHz. As forecast in the weather shown in Figure 7, we suffered from strong winds which unfortunately toppled the 10/24 GHz rig at one point. The setups are shown in Figures 9, 10 & 11. The two groups met for dinner in Phoenix that night, as it was the only chance to get together. Unfortunately Jim K0KFC had to return to Wisconsin due to a family emergency, but was able to leave his 10/24 GHz rig for liaison purposes.

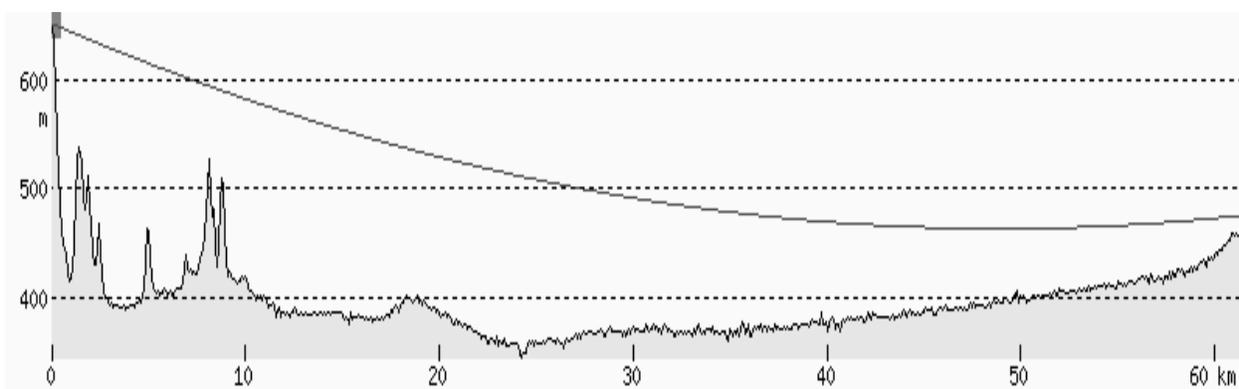


Figure 9 Shaw Butte to San Tan Mountain Park Path Profile



Figure 10 W5LUA's 47/78 GHz & K8ZR's 78 GHz Setup at San Tan Mountain Park



Figure 11 VE4MA with K0KFC's 10/ 24 GHz Liaison Rig at San Tan Mountain Park



Figure 12 N0IO & W7QQ 24, 47 & 78 GHz Setup on Shaw Butte

Operating Results Day #2 Saturday March 9, 2019

The drive from my winter home in Apache Junction, to the summit of Mt Lemmon was nearly 3 hours. We met with Steve KJ7OG, his wife Clare and Ron K7RJ for lunch in Tucson and afterwards proceeded up a fully paved road to the astronomical observatory at the summit where Steve, who is an alumnus of the University of Arizona, had arranged for access. The ground was snow covered but firm and the winds light with a temperature of about 30°F or -1°C (see Figure 12 & 13).



Figure 13 W5LUA, K8ZR, & VE4MA with 78 GHz on Mt Lemmon

The second team again met up with Kevin AD7OI and Tammy KI7GVT and made the journey up the rough road to the White Tanks Mountain site, which with the lower altitude was considerably more comfortable than mount Lemmon with no snow (see Figure 14).

A 10 GHz beacon exists atop White Tanks and this was used to provide initial antenna alignment from Mt Lemmon and after shutting off the beacon, 10 GHz became the means of liaison. 78 GHz signals from N0IO were quickly established at Mt Lemmon, but there was difficulty in the opposite direction. After some chatter to allow system optimization, an intermittent coaxial cable connection was found and solid S7-8 QSOs were concluded. The "trouble shooting" process at White Tanks was videoed and is available on YouTube by searching for 78 GHz W7QQ-VE4MA. This strength of signals is clearly apparent on this video. Notably the smaller 47/78 N0IO rig with Horn antennas was just as strong as the dish rig as it had the same ERP (see Figure 15). In additional QSOs were made by W5LUA, K8ZR, N0IO and W7QQ on 47 GHz.



Figure 14 The View from Mt Lemmon towards White Tanks



Figure 15 Aerial Drone View of N0IO, W7QQ & AD7OI on White Tanks

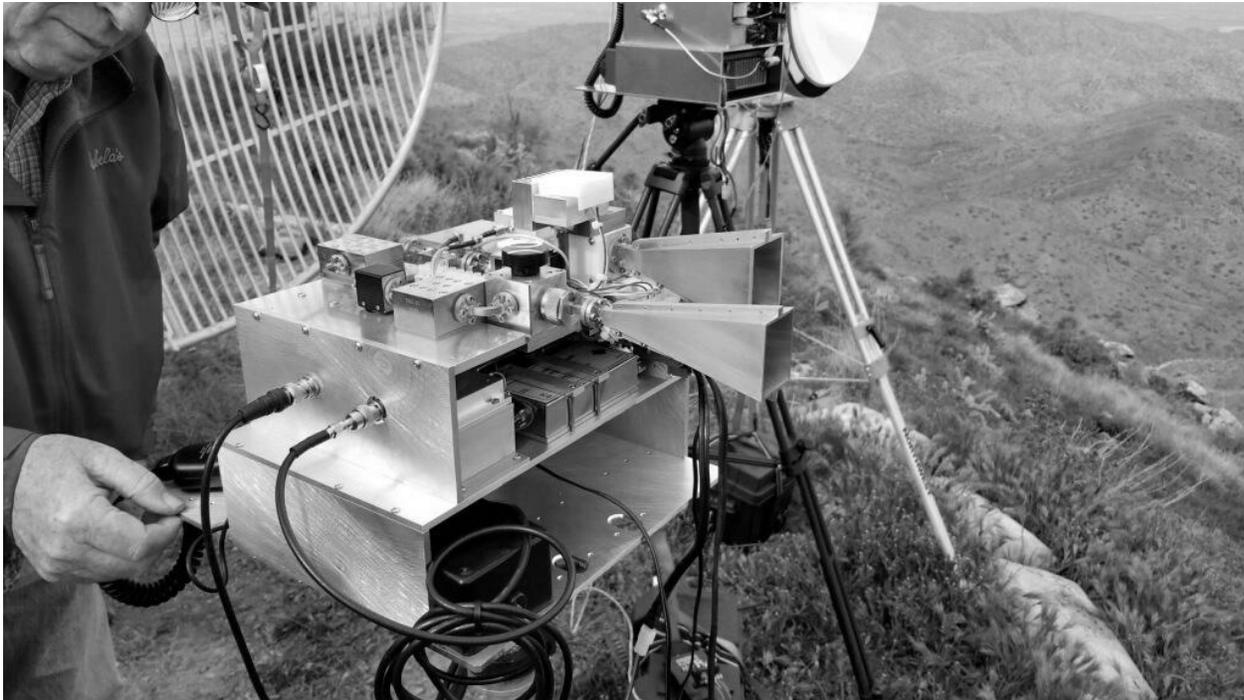


Figure 16 W7QQ with N0IO's Rover Rig for 47 & 78 GHz

Operating Results Day #3 Sunday March 10, 2019

We had originally intended to try the Mt Lemmon to Mt Union 164 mi (265km) path, however Mt Union was not accessible do to over 3ft (1m) of snow that had fallen in the previous week. But buoyed by the outstanding signals on Saturday we were quite optimistic with our chances to make it on the 173 mi (280 km) Mt Lemmon to Mt Harquahala path (see Figure 5). We expected however that the obstruction loss might make it difficult. We assembled in Tucson about 10 am and the weather was marvelous, sunny and about 70°F or 21°C, however as we got within a few thousand feet (1 km) of the summit the weather started to change with cloud rolling in and very high winds causing the pine trees to move violently ! We got set up at the same location; however it took 3 people to mount the 30 in 10/24 liaison dish and 2 people to hold it up! W5LUA and K8ZR set up their equipment downwind of their vehicles. The VE4MA 78 GHz rig was held down with a large cement block and tethered to a guyed street light standard (see Figure 16). Operating the FT817 Radio in the cold wind required headphones and bare hands.

The team at Harquahala drove on another difficult mountain road but did not get any of the bad weather conditions and had a spectacular view (see Figures 17 & 18).

We were able to establish contact on 10 GHz but the signals were very poor. No signals were found on 47 or 78 GHz in spite of numerous attempts! This was truly disappointing and we looked for an explanation. We later obtained the weather data (see Figure 19) and found that the sustained winds on Lemmon were 50 mph (80 km/h) with gusts to 64 mph (104 km/h). More significantly for 47 & 78 GHz was that the Dew point temperature had climbed dramatically to about 30°F or -1°C. This is a significant increase from 0°F or -18°C on Day 2 and would have given an addition loss of ~ 28 dB, which is certainly enough to kill any signals on this path.



Figure 17 Miserable Windy & Cold Operating Conditions on Mt Lemmon



Figure 18 View from Mt Harquahala

Summary

We certainly were able to show that Arizona can provide some excellent mmWave paths when the weather cooperates! We look forward to going back to test Mt Lemmon to Mt Union and further trying to break the 289 km world record. We also want to test on 122 & 241 GHz and optical frequencies.

On behalf of the team members I would like to extend our appreciation to Steve KJ7OG, his wife Clare, Ron K7RJ, Kevin AD7OI, Tammy KI7GVT and Jim K0KFC for their support in planning, executing and documenting this adventure.



Figure 19 Second Group Operating at Mt Harquahala

Weather History Graph
March 10, 2019

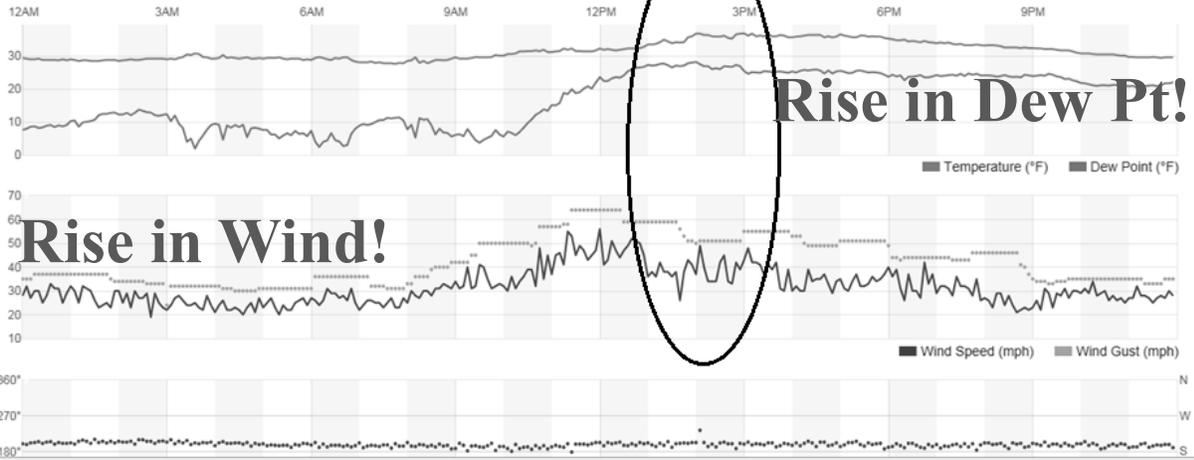


Figure 20 Mt Lemmon Weather Information for Day 3 Tests

Waveguide Cross-guide Directional Couplers

Paul Wade W1GHZ ©2019

www.w1ghz.org

When my new 47 GHz transverter arrived from DB6NT¹, the first goal was to put it on the air. I had no test equipment and couldn't find any of the right size WR-19 waveguide in the junk box. Lacking a waveguide switch for TR switching, I decided to make two separate horns and get it on the air.

At our local makerspace, The FoundryVT², I made a pair of Skobolev dual-mode horns³ using a CNC lathe. The horns have an estimated gain of 23 dBi. I put them on the transverter breadboard, shown in Figure 1, and made some contacts. Best DX was 90 km, beyond line-of-sight over the ocean between the islands of Block Island, RI, and Martha's Vineyard, MA. In the 2019 10 GHz & Up contest, I increased it to 123 km, line-of-sight between Mt. Mansfield, VT and Mt. Washington, NH.

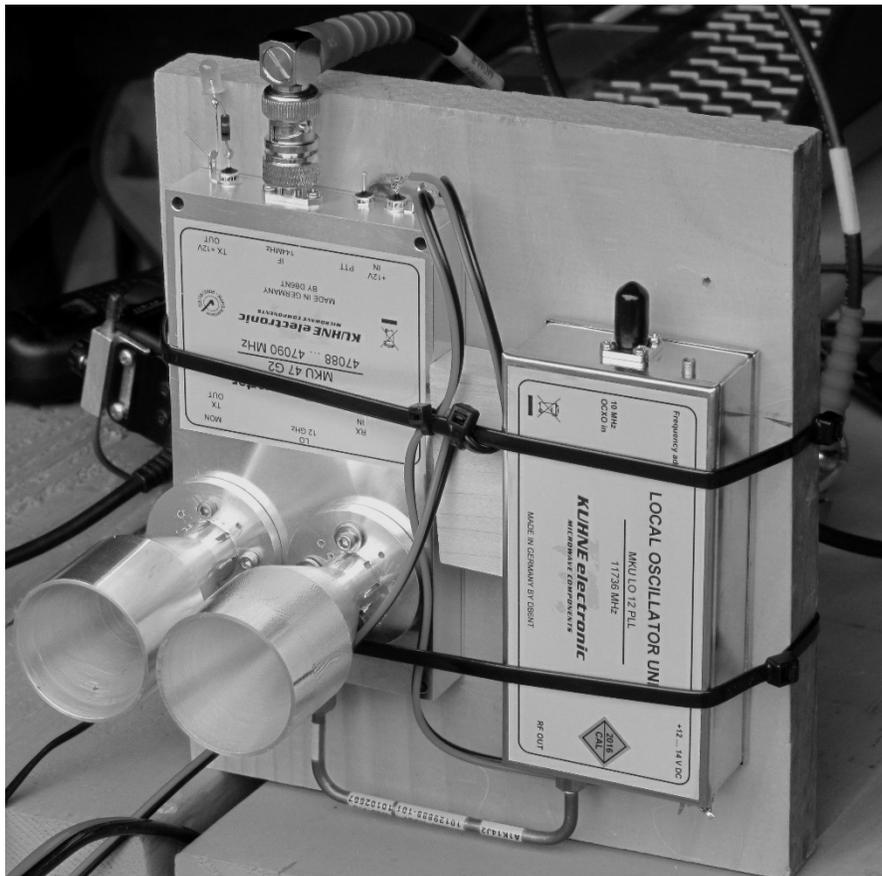


Figure 1 – 47 GHz DB6NT Transverter breadboard with separate horns

With the CNC machinery at the makerspace, I have been able to make a number of waveguide components in WR-19 and other sizes, including the OE9PMJ filters I described⁴ at Microwave Update 2018. However, I have only been able to tune and test them at VHF conferences where a fancy VNA is available. Thanks to the suppliers of this equipment for making it available.

Waveguide Directional Couplers

I have found and used surplus waveguide directional couplers for 10 and 24 GHz and other frequencies, but never for 47 GHz. Could I make them for 47 GHz? I have seen three types: two with parallel waveguides, both broadwall and sidewall, and the other with crossed perpendicular waveguides; versions of all three types are shown in Figure 2. The parallel variety requires bends in one waveguide to access all four ports, an additional complication, so I decided on the cross-guide version, which could be machined in a block of aluminum.

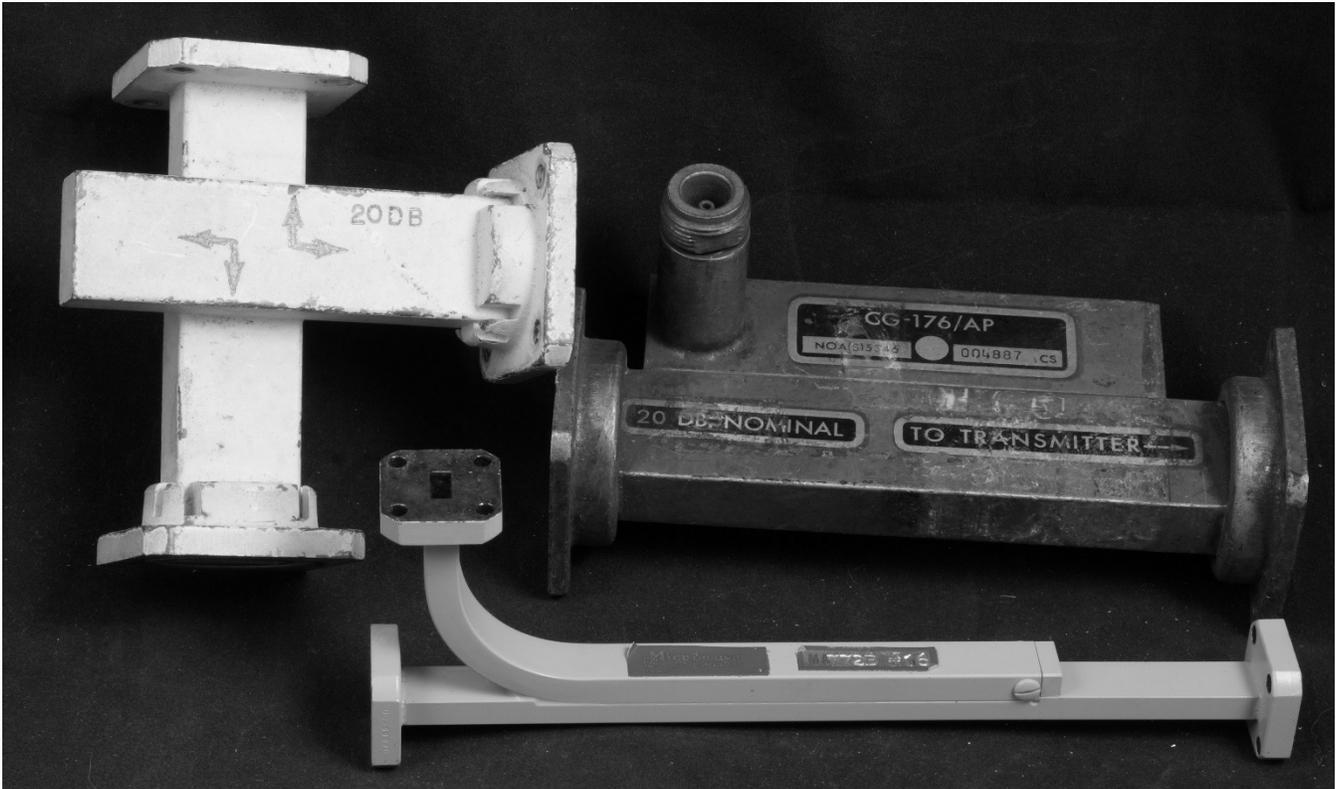


Figure 2 – Surplus waveguide directional couplers. Clockwise from left: cross-guide coupler in WR-75, sidewall coupler in WR-90, and broadwall coupler in WR-28.

The WWII RadLab series has 44 pages on directional couplers⁵ in Volume 11, but most of them look rather difficult to make. Publications on waveguide have appeared frequently since then, but details for waveguide directional couplers are scarce. I found bibliographies^{6,7,8,9} of all directional coupler papers between 1938 and 1966, but very little with useful details.

Two types of cross-guide couplers have been described – the Moreno type with two "X" shaped holes between the crossed waveguides¹⁰, sketched in Figure 3, and one with three round holes between the crossed guides. The cross-guide coupler is of this type. The X-shaped holes are reasonably large at 10 GHz, but the dimensions are very small at 47 GHz, beyond my machining capabilities, while round holes are easy to make in any size.

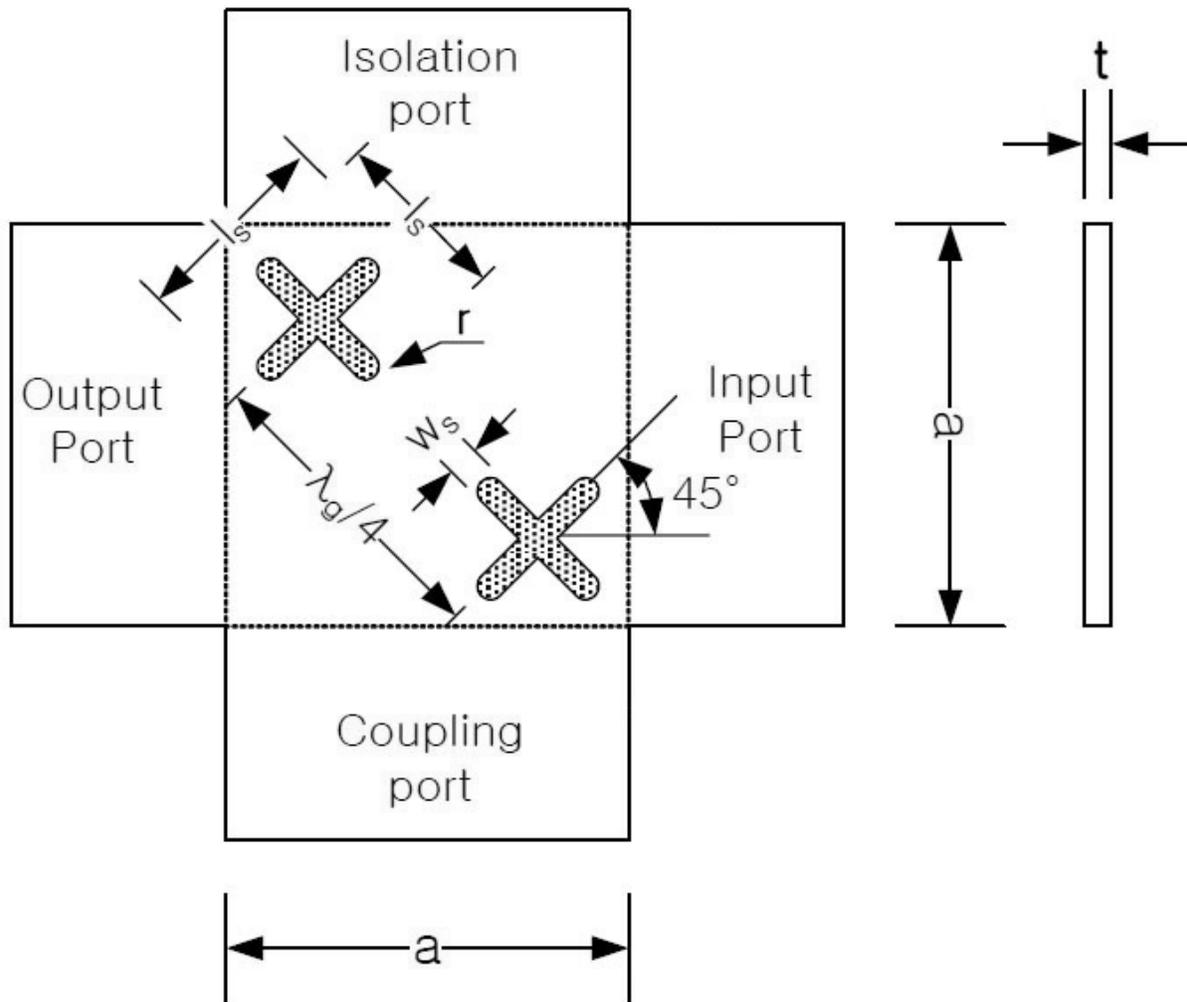


Figure 3 – Moreno cross-guide directional coupler sketch

The only publication I found for the round hole type is a single graph labelled “ROUND HOLE CROSS GUIDE DIRECTIONAL COUPLER” in the *Microwave Engineers Handbook*¹¹ attributed to Gershon J. Wheeler, showing two diagonally-placed holes plus a third, smaller, hole forming a triangle, with all holes $\lambda g/8$ from the centerlines of both guides. The graph shows hole size vs coupling for couplings between 21 and 42.5 dB for the larger holes; the smaller hole is specified as 2/3 of this diameter.

I later found that Wheeler had patented¹² the three hole directional coupler, with the patent assigned to Raytheon. Figure 4 is a sketch of this coupler from the patent. The patent has expired, so we aren't going to infringe. The patent has lots of claims, but little useful information: “proven by empirical methods to increase directivity.” One claim is that a third hole can improve the version shown in Figure 2 as well. Of course, the real details would have been proprietary information.

CROSS GUIDE DIRECTIONAL COUPLERS

Filed Dec. 15, 1954

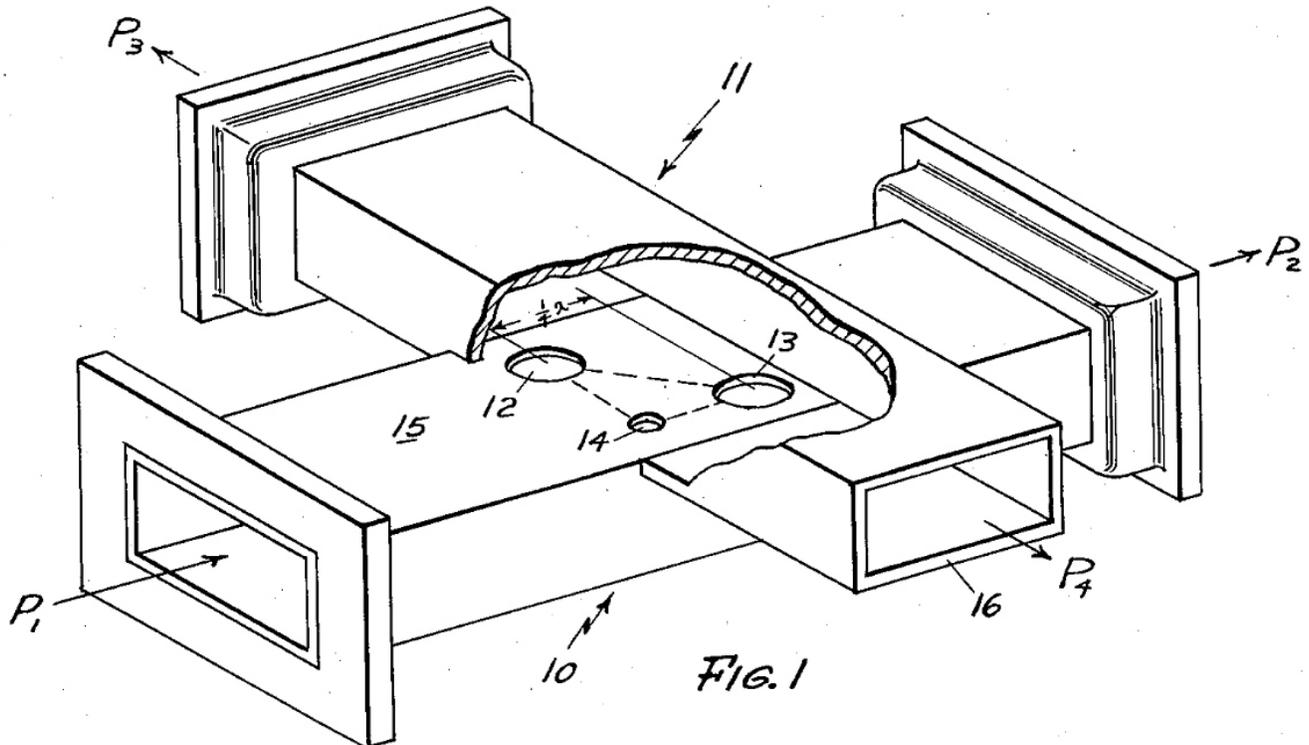


Figure 4 – Round hole cross-guide directional coupler sketch

What wasn't mentioned in the graph is the thickness of the wall between the crossed waveguides that the holes pass through. The patent says the holes are in the wall of one waveguide, with the wall of the other cut away. Since I machine both guides out of a single block of aluminum, leaving the wall between them, 0.5 mm seemed like a comfortable thickness.

The graph seemed like a good starting point. Since I don't have any 47 GHz test equipment, starting with a WR-42 coupler at 24 GHz seemed like a good way to prove the concept, with the wall thickness scaled to 1 mm. I did an HFSS¹¹ simulation of a 30 dB WR-42 coupler with the published dimensions – the results, shown in Figure 5, were not too promising, with only about 12 dB directivity. This result plus the round numbers for hole locations suggest that the graph is only a first order calculation, probably based on the proverbial "thin" wall. However, simulations with thinner walls only improved directivity a couple of dB, but the coupling increased with thinner walls, bracketing the 30 dB target, as shown in Figure 5.

**Three-hole Cross-guide Coupler
WR42
Coupling and Directivity
Handbook Dimensions**

W1GHZ

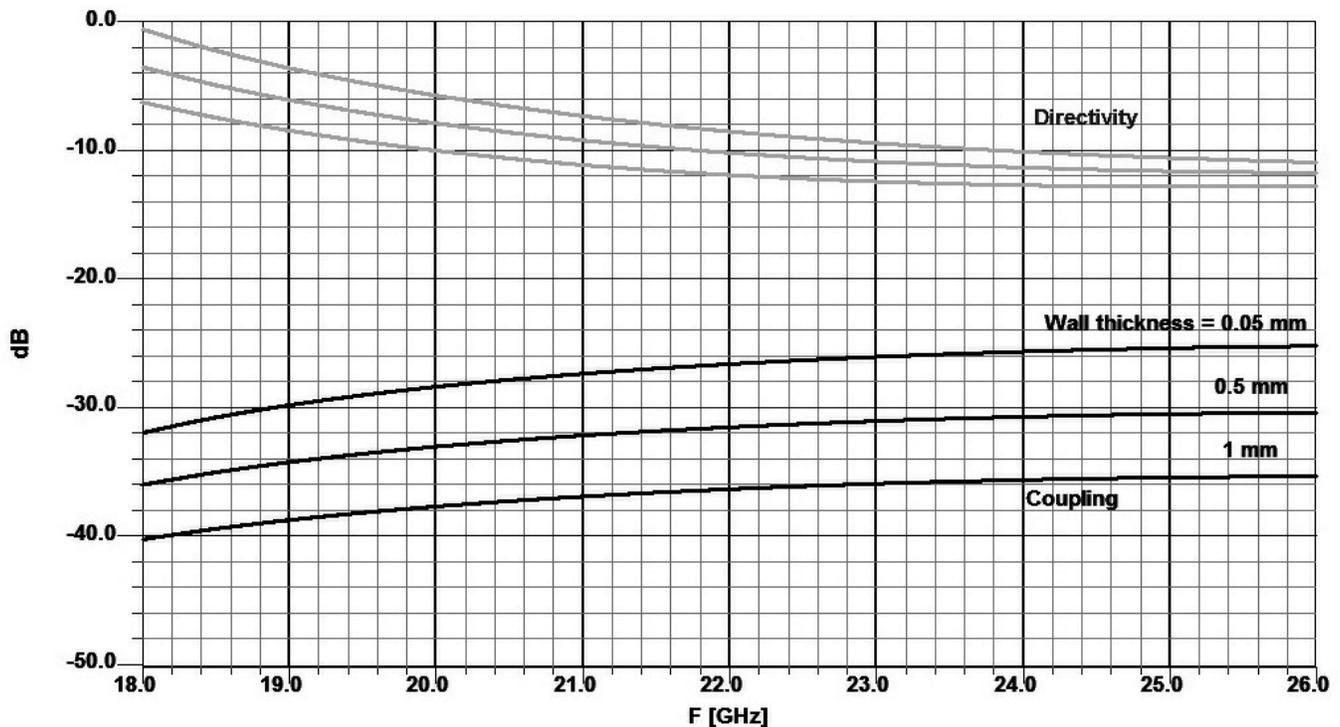


Figure 5 – Simulation of 30 dB round hole cross-guide directional coupler with dimensions from Wheeler graph

Directivity

A directional coupler with only 12 dB of directivity isn't very useful. Directivity is the difference between coupled forward power and coupled reflected power with a perfect termination; in other words, leakage into the reflected port. This means that the indicated Return Loss with a perfect load would be 12 dB. Any reflected power from different terminations would add or subtract from the leakage, depending on phase, giving bogus readings. Forward power indication also includes leakage, so that could vary by +/-6% (-12 dB) from true power. Note that hams bragging about -35 dB return loss (or 1.036 VSWR) had better know the directivity of their test equipment.

Could the directivity of the cross-guide couplers be improved by adjusting the holes? Some trial and error suggested that the center frequency of the coupler was too high, so the holes should be further from the center line. Directivity improved by making the smaller hole a bit larger, but only over a relatively narrow bandwidth. Coupling is fairly flat over a wider bandwidth, but not good directivity. Fine tuning of hole placement and size produced 30 dB and 40 dB couplers in WR-42 with flat coupling and excellent directivity from 24.0 to 24.5 GHz. An attempt at increasing the hole diameters to produce a 20 dB coupler was not successful – large holes that graze the waveguide wall only resulted in 23 dB coupling.

Final dimensions after simulation suggest that Wheeler's graph for coupling was reasonably close, but the other dimensions are not as good. The hole distance from guide centerlines is closer to $\lambda g/6$ rather than $\lambda g/8$, and the small hole diameter providing best directivity is around $0.8 D$ rather than $2/3 D$. Actual dimensions of couplers that I made and measured are shown in Table 1.

24 GHz

I machined both 30 dB and 40 dB WR-42 cross-guide couplers on a CNC milling machine. Hole location is controlled by the CNC using the same indexing as the waveguide cuts, and hole diameter is tested and fine-tuned using pin gauges. Figure 6 is a photo of one side with waveguide and holes, and an assembled coupler. Twenty tapped holes are required for assembly and waveguide flanges – all the holes would never line up without CNC. An additional complication is that the threaded holes intersect with the joint between metal blocks.

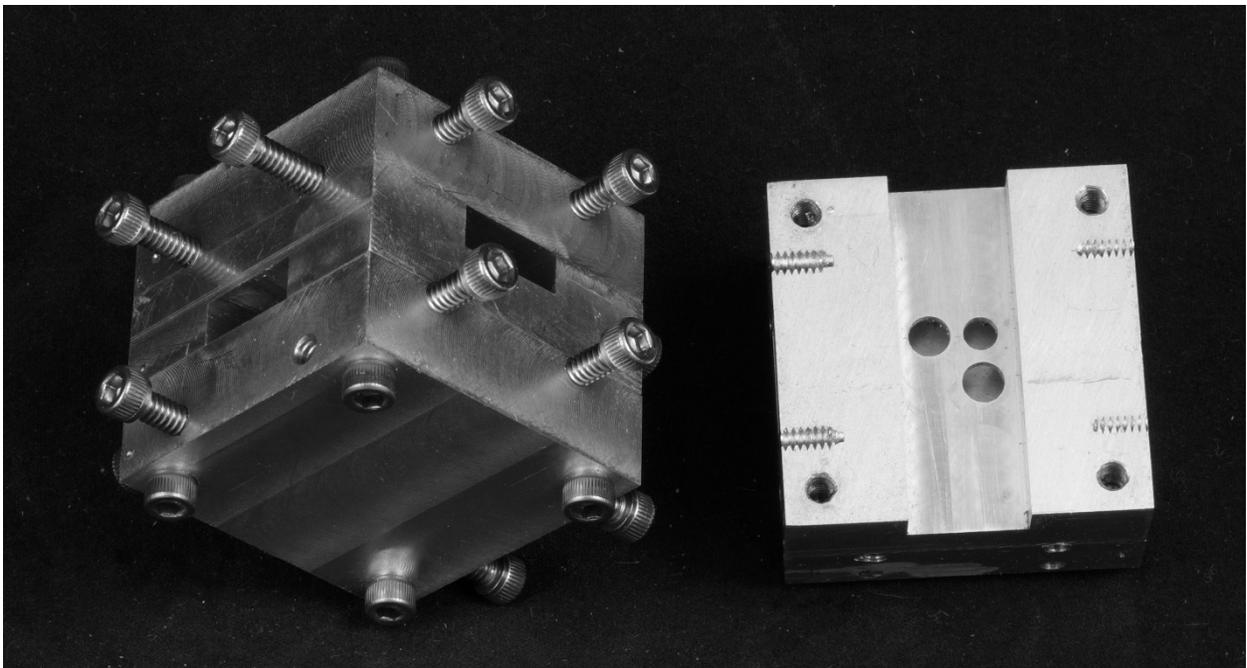


Figure 6 – Machined round hole cross-guide coupler for 24 GHz in WR-42 waveguide

I got a chance to test one of the finished couplers at Microwave Update 2017 on an Anritsu VNA with full waveguide calibration; coax transitions add uncertainty and usually degrade directivity. Thanks for to Anritsu for providing this equipment and to Jeffrey Pawlan, WA6KBL, for the waveguide calibration. The results, shown in Figure 7 are very good, with 35+ dB directivity from 24.0 to 24.5 GHz. Coupling is 38.3 dB, slightly less than the target 40 dB.

The measured results are compared with simulation in Figure 8. The measured coupling is slightly stronger than simulated, and the measured directivity is best at 24 to 24.5 GHz while the simulated directivity is best at a bit higher frequency – a fortunate difference.

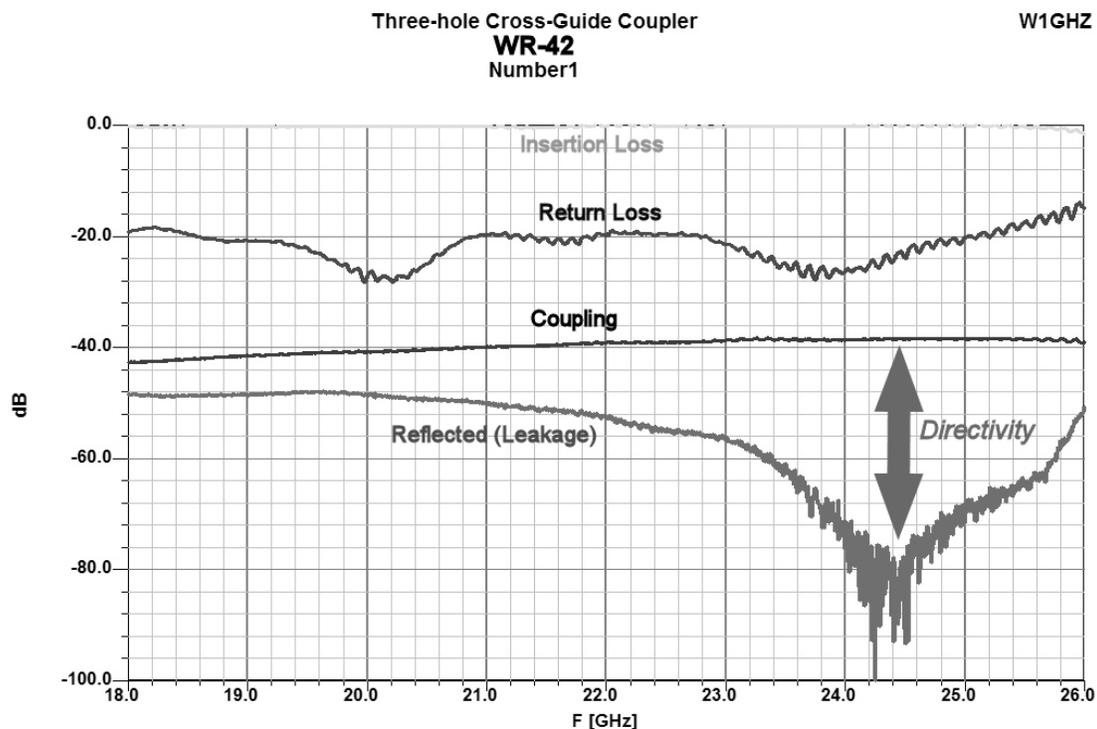


Figure 7 – Measured performance of round hole cross-guide 40 dB directional coupler for 24 GHz in WR-42 waveguide

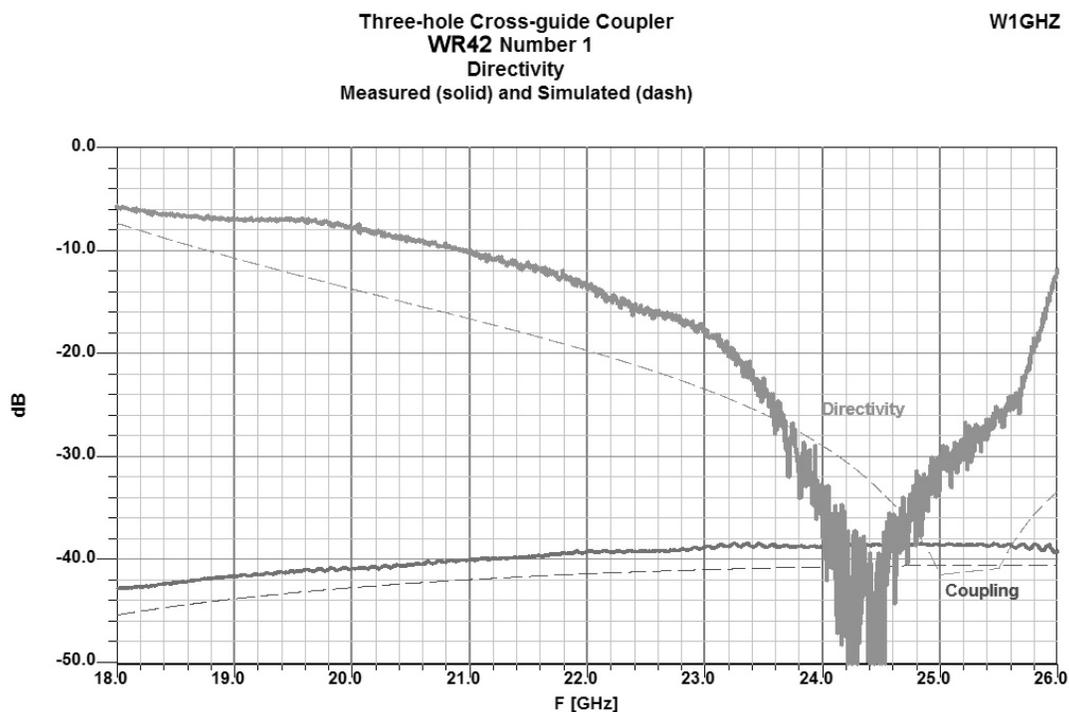


Figure 8 - Measured performance comparison to simulation of round hole cross-guide 40 dB directional coupler for 24 GHz in WR-42 waveguide

The second WR-42 coupler, targeted at 30 dB coupling, was measured in coax with homebrew transitions on a Rohde & Schwarz ZVA67 VNA at the 2018 Eastern VHF/UHF Conference thanks to Greg Bonaguide, WA1VUG. Results were not as good – the homebrew transitions appear to have significant mismatch which compromised the results. Coupling was about -29 dB, slightly stronger than the target.

47 GHz

Encouraged by the 24 GHz results, I machined three WR-19 cross-guide couplers, two with 30 dB nominal coupling and one with 40 dB nominal coupling. I was only able to measure these couplers in coax using WR-19 to SMA transitions that I made, at Microwave Update 2018. SMA connectors are way beyond their specified range at 47 GHz, and a pair of transitions show about 2 dB loss, adding uncertainty to the coupling. The results for one of the 30 dB couplers, shown in Figure 8, are not great, but the coupling is close to the 30 dB target. The other two couplers showed similar results.

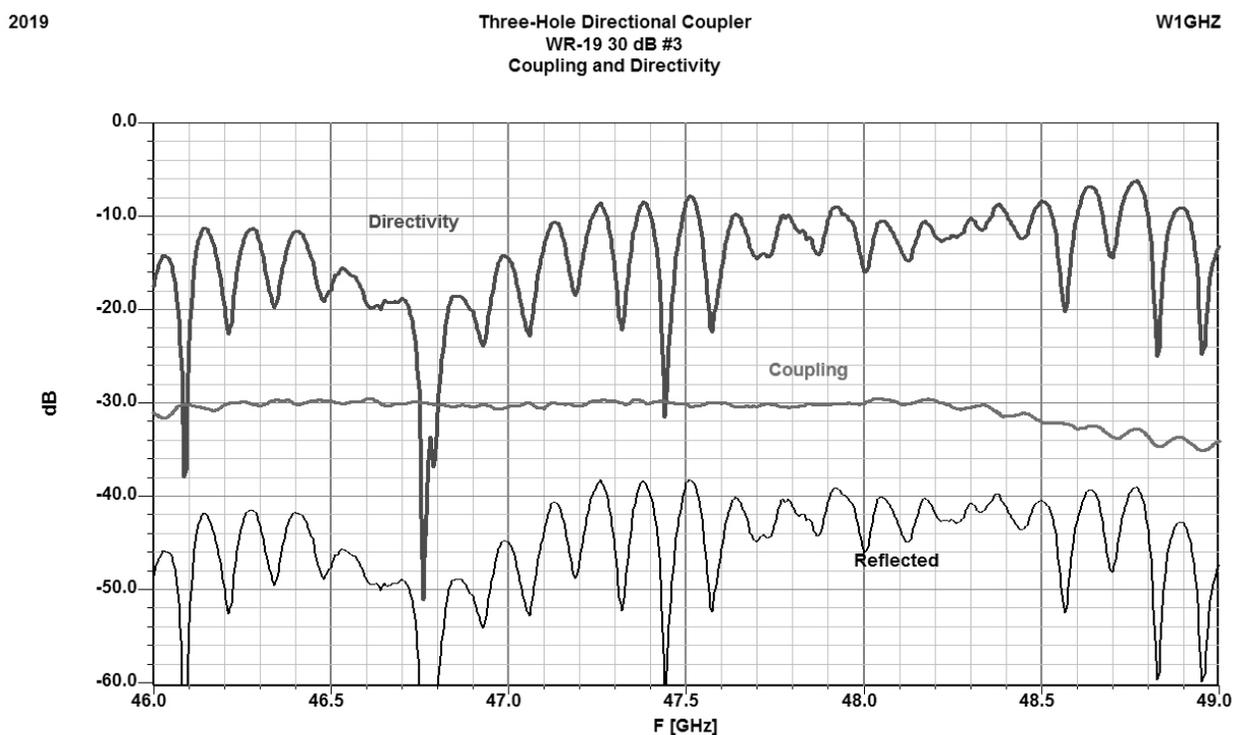


Figure 9 - Measured performance of round hole cross-guide 30 dB directional coupler for 47 GHz in WR-19 waveguide

I am not ready to buy a 47 GHz amplifier from DB6NT, and my transverter output is only 30 or 40 milliwatts, so the 30 and 40 dB couplers aren't much use yet. A 20dB coupler would be useful, but the strongest possible coupling in simulation is about 23 dB. However, all the measured couplers have slightly stronger coupling than expected, and making the wall between the waveguides thinner also increases the coupling. Making these changes might get close to 20 dB coupling.

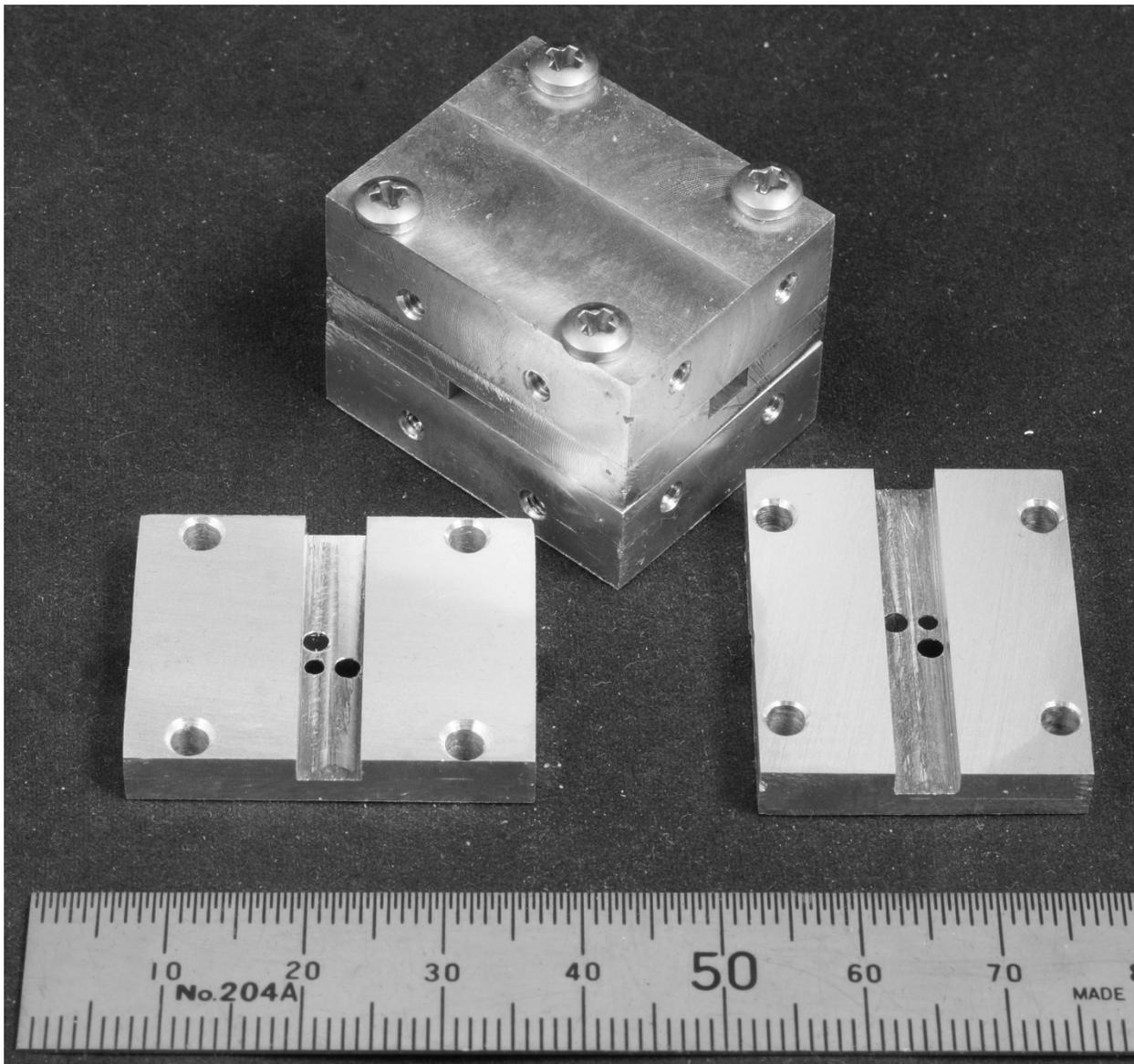


Figure 10 - Machined round hole cross-guide 20 dB directional coupler for 47 GHz in WR-19 waveguide

I machined two more WR-19 couplers with large holes and a thin wall, which ended up somewhere between 5 and 10 mils thick – WR-19 waveguide is too small to fit a micrometer inside. The large holes may be seen in Figure 10. Measurements were made at the 2019 Super VHF Conference, with results shown in Figure 11. Both have about 22 dB coupling including coax transitions; a pair of transitions alone has about 2 dB loss, so the actual coupling may be close to 20 dB. Directivity is very good, at least 30 dB from 47 to 47.5 GHz. These couplers are actually useful for measuring Return Loss with an HP432 power meter.

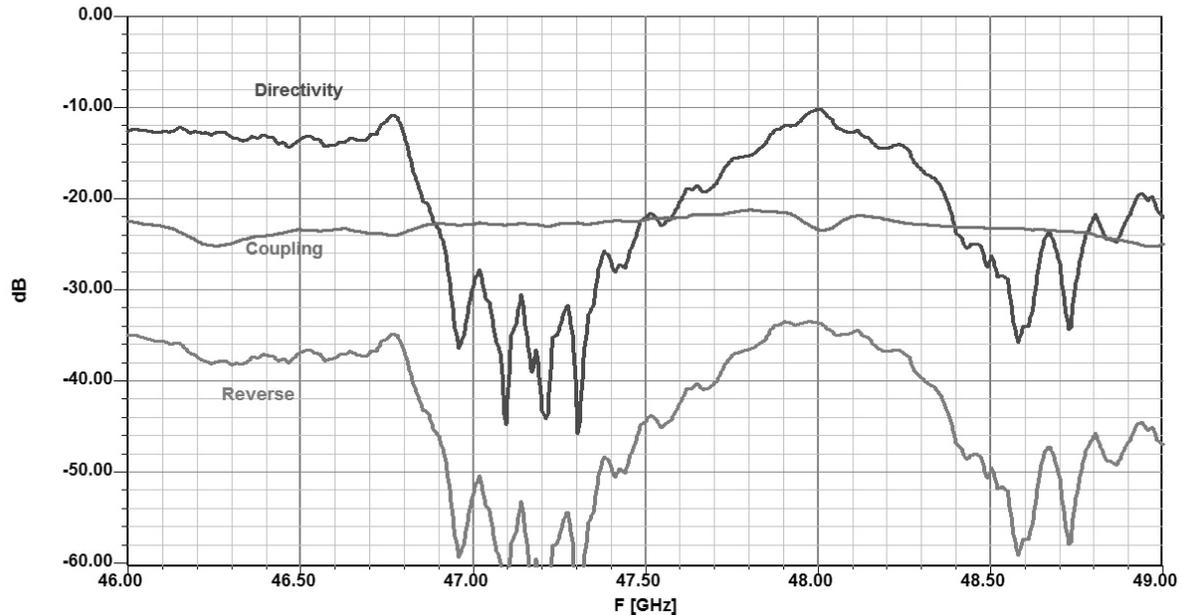


Figure 11 - Measured performance of round hole cross-guide 20 dB directional coupler for 47 GHz in WR-19 waveguide

Construction

I machined the 24 and 47 GHz directional couplers as three aluminum blocks: a center section with both waveguides, separated by the wall containing the coupling holes. The waveguides are simply rectangular grooves in the center block, with flat top and bottom pieces forming the fourth wall of the waveguides. The two perpendicular waveguides are offset vertically by the wall thickness, so the screw holes for the waveguide flanges are asymmetric. Tapping the threads for 4 flanges plus assembly screws, a total 20 threaded holes, is rather tedious.

A possibly better way to produce these directional couplers would be to photo etch the holes into a thin metal plate, to be placed between two metal blocks. Each block would have a waveguide machined into it, with opposite sides of the metal plate forming the fourth wall of both waveguides. Alignment pins would control placement of the coupling holes. Companies offering precision photo etching have displays at microwave trade shows.

10 GHz

Some serious EME stations run real power at 10 GHz, and waveguide is required – SMA connectors aren't good for much more than 50 watts. Directional couplers for 10 GHz, in WR-90 waveguide, are occasionally found in surplus. A quick check of ebay finds a number of 10 and 20 dB couplers, but nothing at 30 dB or higher. For instance, at least 40 dB is needed for 100 watts. I worked out hole dimensions for both 40 and 50 dB of coupling in WR-90.

Rather than machine large blocks of aluminum, I used surplus pieces of ordinary WR-90 waveguide with the holes drilled in one piece and a window cut out in the other, as shown in Figure 12, then soldered them together. To make alignment easier, the wall with the holes was notched slightly to hold the guide with the window in place. As a result, the wall was thinned slightly, to about 1mm. I made the holes, notch, and window using CNC so they fit perfectly, but WR-90 is large enough that a coupler could be built with hand tools with some care.

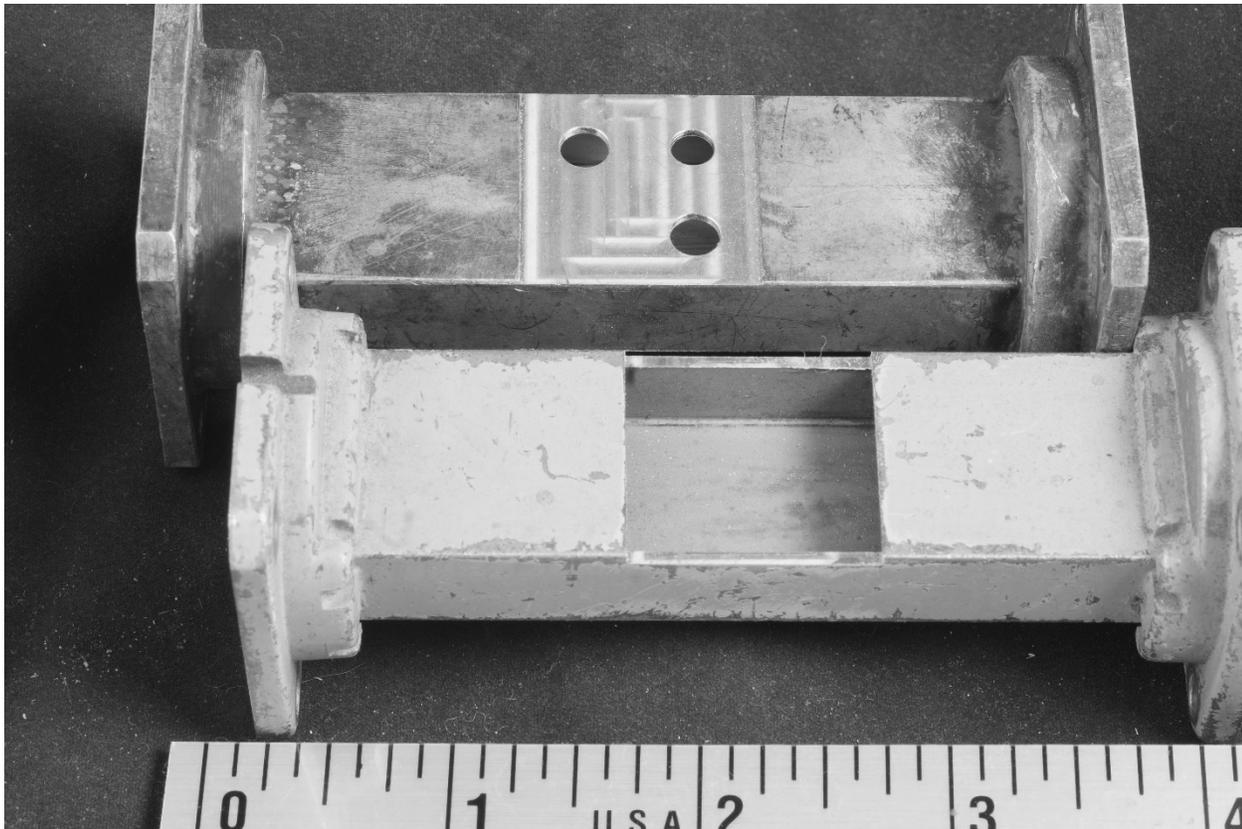


Figure 12 - Round hole cross-guide directional coupler for 10 GHz in WR-90 waveguide fabrication from surplus waveguide before soldering together.

The couplers, shown complete in Figure 13, were measured at the 2019 Super VHF Conference. Measuring the couplers required waveguide to coax transitions for the VNA. I don't have enough professional transitions, so some homebrew ones were used as well. These are good enough for ham rigs, but not for precision measurements with the >80 dB range required for a 50 dB directional coupler.

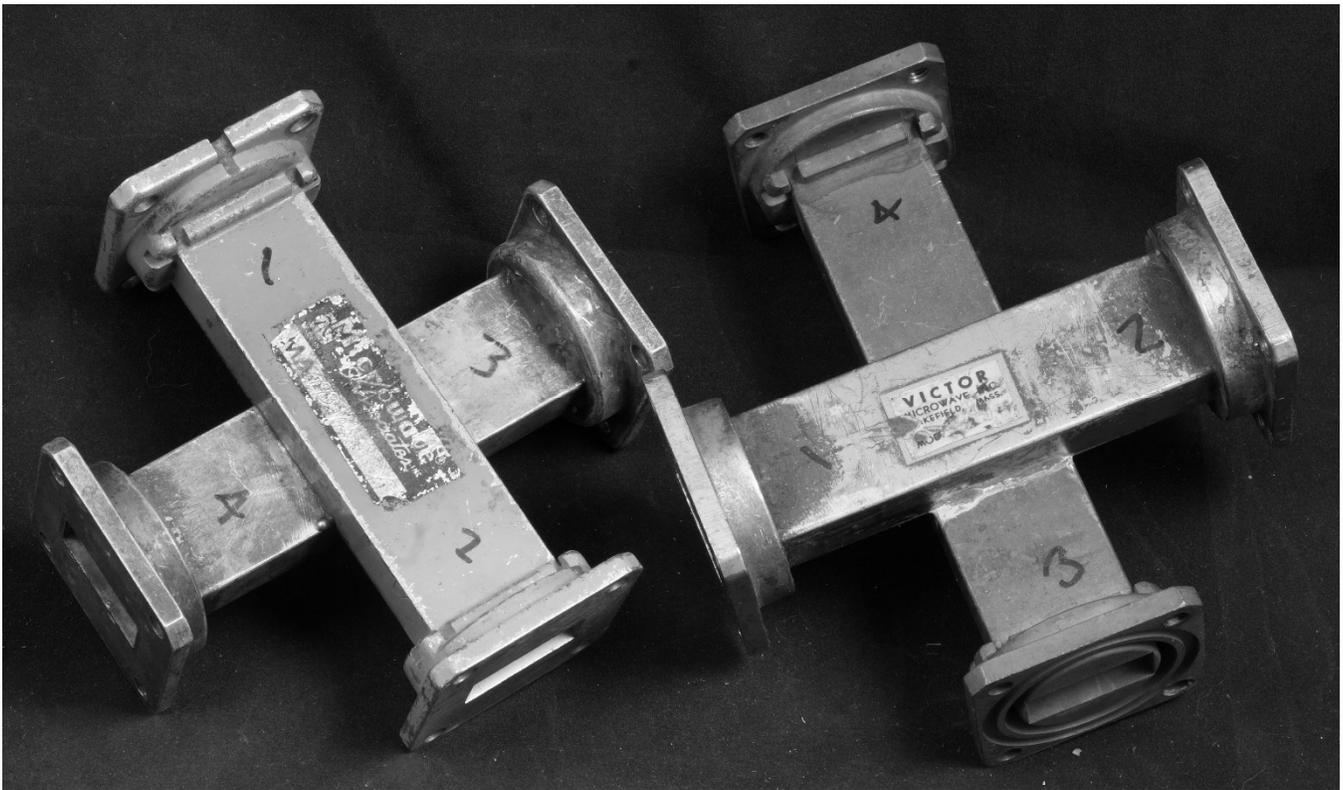


Figure 13 - Round hole cross-guide directional couplers for 10 GHz in WR-90 waveguide

Measurement results are not completely available – I apparently did something wrong in saving the S-parameters, so I only have screenshots of the plots. Like the higher frequency couplers, coupling at 10.368 GHz is slightly stronger than predicted by simulation: 37.3 dB for the nominal 40 dB coupler, and 46.3 dB for the 50 dB one. Directivity is not as good as expected, ~20 dB for the nominal 40 dB coupler, and ~16 dB for the 50dB one; this is probably limited by the coax transitions. These couplers are useful for measuring QRO power: with a power meter like the HP432, limited to 10 milliwatts, — 46.3 dB coupling allows measurement of up to 426 watts.

Summary

Waveguide directional couplers with good directivity were made for 47, 24 and 10 GHz. These couplers can be used to measure high power directly in a waveguide system, and as instrumentation to measure Return Loss. Hole dimensions for all of these directional couplers are shown in Table 1, to enable others to replicate the couplers. Not included, but available on request, are some other simulated dimensions.

I do not plan to produce and sell any of these directional couplers, but would be willing to assist any ham who wishes to produce his or her own or to make them for sale.

Acknowledgements

Very little of this machining would have been possible without the assistance and mentoring of Tom Bishop of The FoundryVT and other members of this fine Makerspace.

Measurements of these directional couplers made possible by the availability of test equipment at conferences from Rohde & Schwarz, courtesy of Greg Bonaguide, WA1VUG; Agilent; and Anritsu, with assistance from Jeffrey Pawlan, WA6KBL.

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Three-hole Cross-guide Directional Coupler Dimensions

W1GHZ
2019

<u>Waveguide</u>	<u>Coupling</u> Target dB	<u>Coupling</u> actual dB	<u>Wall</u> thickness inches	<u>Large</u> <u>hole</u> Diameter inches	<u>Small</u> <u>hole</u> Diameter inches	<u>off CL</u> distance inches
WR-42	30	29.7	0.045	0.146	0.120	0.089
	40	38.3	0.045	0.122	0.098	0.089
WR-19	21.8	~21	~.010	0.079	0.062	0.050
	30.4	~30	0.029	0.078	0.062	0.0484
	40	~39	0.029	0.062	0.050	0.0484
WR-90	40	37.3	0.04	0.209	0.184	0.228
	50	46.3	0.04	0.158	0.138	0.225

Table 1

Arduino Controllers for the Microwave Operator

Greg McIntire, AA5C

Arduino and Raspberry PI microcontrollers have become popular and found widespread use in amateur radio applications. Over the past few years it seems almost every issue of QST has an article on an Arduino application, with good reason. They are versatile and low-cost with a variety of low-cost, easy to interface hardware add-ons and a wealth of open-source software available. I became interested in Arduino controllers as a means to control phase locked loops (PLL) and direct digital synthesizers (DDS) that use a digital control interface and have experimented with a number of PLL and DDS using Arduino controllers. I reported on the performance of several low-cost PLL evaluation boards based on the Analog Devices ADF4351 and ADF5355 PLLs, as well as other projects, in a 2017 Microwave Update article (Reference 1). This 2019 article reports on the results of several Arduino-based projects since that time and includes general information about the controllers and available supporting hardware intended to help you apply them to your needs.

Most modern PLLs use a serial peripheral interface (SPI) bus for control and status. This makes control options robust but also requires some sort of controller in order to use the part since a number of registers need to be loaded in order to make the parts function. PLL and DDS register lengths are now generally 32-bits which make for easy code reuse. PLLs with other length registers can be controlled as well. I had an older (circa 1996) National LMX2316 PLL evaluation board that I wanted to get running. Three 21-bit registers need to be loaded for it to operate. Coding turned out to be simple since the first 11 bits of each 32-bit register transfer are serially shifted into the “bit bucket” and the last 21-bits containing the necessary control bits end up in the correct position. The hardware interface was also the simple four-wire SPI interface, and since the LM2316XL uses 5 volt logic, no level shifting resistors were required to interface it to the Arduino Uno. This particular evaluation board is older and has a limited frequency range compared to current devices, but the project illustrates how an Arduino can be used on a range of different PLLs (Figure 1 LMX2316 Evaluation Board Controlled by an Arduino Uno).

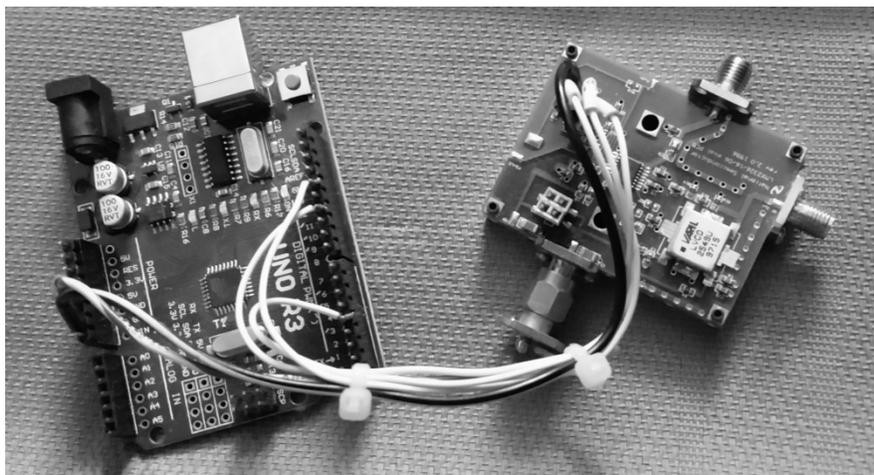


Figure 1 LMX2316 Evaluation Board Controlled by an Arduino Uno

Studying the register set of each particular PLL is the best way to understand the capabilities and limitations of each device. Current PLLs have many features you can exploit if you are willing to dig in to really understand the device. For instance, they can be programmed for best phase noise performance or best spur performance. Phase can be controlled on some to support the generation of complex waveforms. Some device manufacturers make software tools available that generate the register values for the particular frequencies and configurations you select. I'm familiar with several of the Analog Devices tools and they can greatly simplify the generation of the control register values. However, I still go through the data sheets in detail to better understand the part. The software tools are generally useful for deriving control values for a single frequency. When you are changing frequencies on the fly in real-time, what I refer to as dynamic control, you need to understand the frequency control formulas sufficiently in order to program the controller. The same applies for other real-time controls like on-off keying of the RF output.

Some of the PLLs are programmable with frequency resolution down to a Hertz or so. These parts can tax or exceed the capabilities of some of the simpler controllers. The ADF5355 is a fractional-N PLL with the following frequency control register lengths that allow near Hz frequency resolution: INT 16-bits, FRAC1 24-bits, FRAC2 14-bits, and MOD2 14-bits. The length of these frequency control registers should give you an idea of the achievable resolution.

I've found that the simple 8-bit Arduino Uno works well for loading 32-bit register values in fixed-frequency LO applications where values are pre-calculated. My experience in calculating the values in real-time was that I could get the precision needed for the ADF4351 using an 8-bit Arduino Uno, but not for the ADF5355, even using extended length arithmetic. A 32-bit Arduino Due did the job nicely and my Arduino Due/ADF5355 combination has been a useful lab tool. I've had some brief email exchanges with Alain, F1CJN, and Ed, W7GLF and they may have worked out a means of using the Arduino Uno for dynamic control of the ADF5355 but I have not seen their code or technique to try it.

Recently I completed upgrading the six transverters in my microwave station (902 through 10368 MHz) with local oscillators (LO) locked to a GPS-disciplined 10 MHz reference. I started with Apollo 32 boards for 758 MHz, 1152 MHz, and 2160 MHz LOs. This was before



Figure 2 AA5C Microwave Transverters

I became aware of the ADF4351 evaluation boards. I then used an ADF4351 board for 3312 MHz and ADF5355 boards for 5616 MHz and 10224 MHz. All of the ADF4351 and ADF5355 boards are

controlled by Arduino Unos. The BG7TBL GPS-disciplined 10 MHz oscillator (green faced box at the top of the picture) feeds a two-way power splitter which then feeds two 4-way splitter/amplifier/filter boards from Down East Microwave. A total of eight outputs are available. The boards are in the chassis shown next to the power meter at the top of the picture. Now each of the six microwave transverters in the shack and a microwave frequency counter are locked to GPS. Not that long ago we would sometimes have to tune up to +/-50 KHz on the IF rig to find the station we were trying to work. That operating variable is now significantly reduced or eliminated, depending on what the other station is using.

A single controller can be used to control multiple PLLs or devices that use a SPI bus interface. An additional I/O pin needs to be assigned for each extra device connected to the SPI bus for the load enable line (LE) and called out separately in the code. I recently completed a 47 GHz transverter that uses two ADF5355 evaluation boards for the LOs. It uses a double conversion frequency plan. The two ADF5355 evaluation boards are controlled by one Arduino Uno (Figure 3 AA5C 47 GHz Transverter) as shown in the lower left of the figure. The additional amplifiers needed to bring the relatively low (~ -8 dBm) output power levels of the ADF5355 evaluation boards (two 12 GHz amplifiers are required for the 12,978 MHz LO and one for the 8.01 GHz LO) can also be seen in the picture. The 32 MHz reference signal is supplied to each synthesizer from a two-way power splitter.

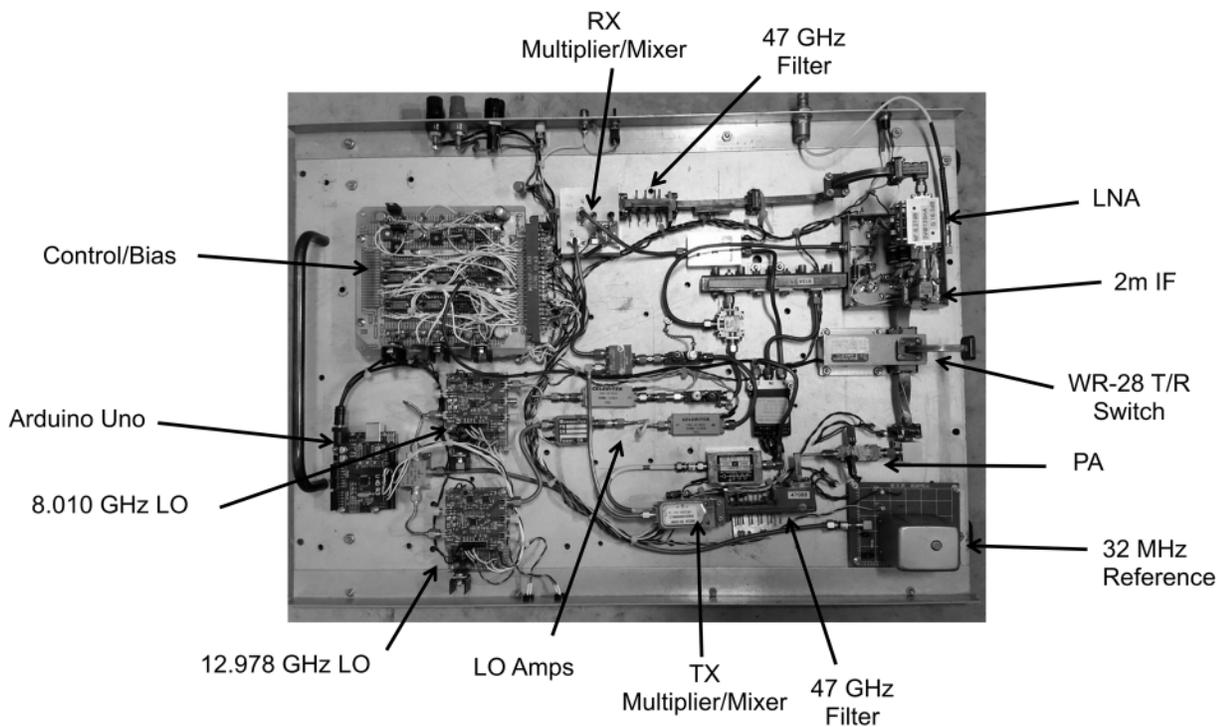


Figure 3 AA5C 47 GHz Transverter

The versatility of using a controller for the PLL is exemplified by the fact that I started the project with a 10 MHz reference. I couldn't calibrate the 10 MHz reference to precisely 10 MHz so I switched to a 32 MHz reference that I could calibrate precisely. A quick recalculation of the register values and flashing them into the controller was all that was required to use the new reference.

The frequency accuracy of the 47 GHz transverter has turned out to be quite good. It is not locked to a GPS-disciplined oscillator but I did calibrate the 32 MHz reference using a microwave counter locked to a GPS-disciplined oscillator. When working AI, W5LUA on 47 GHz, we have generally been within a KHz or two of each other which is pretty amazing at 47 GHz.

I've experimented with the four Arduino controllers shown in Figure 4 Example Arduino/WEMOS Controllers . Clockwise, starting in the upper left hand corner of the figure and working clockwise, these are the Arduino Mega 2560, the Arduino Uno, the WEMOS D1 Mini, and the Arduino Due

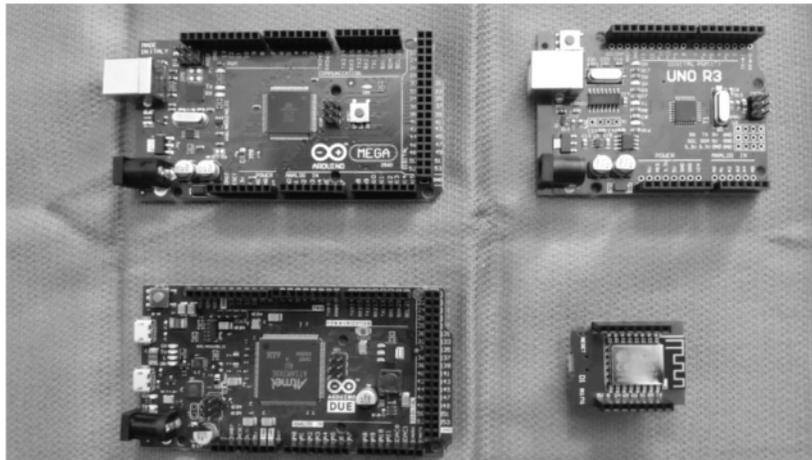


Figure 4 Example Arduino/WEMOS Controllers

Some of the key features of these controllers are tabulated in Table 1 Comparison of Controller Key Features. I extracted these values from the Arduino and WEMOS websites and believe they are accurate but encourage you to check for yourself. Hopefully this will give you a starting place for picking a particular controller for your application. The more expensive ones were around \$20. The Uno and D1 Mini can be found for \$5 or less.

Board	Mega 2560	Uno	Due	D1 Mini
Processor	ATmega2560	ATmega328P	ATSAM3X8E	ESP8266
Clock Speed (MHz)	16	16*	84	80/160
Word Length (bits)	32	8	32	32
I/O Logic Voltage	3.3	5	3.3	3.3
Analog I/O (in/out)	16/0	6/0	12/2	1/0
Digital (I/O or PWM)	54/15	14/6	54/12	11
EEPROM (KB)	4	1	0	512 bytes
SRAM (KB)	8	2	96	4 MB
Flash (KB)	256	32	512	4 MB
SPI Bus Interfaces	1	1	2	1
USB Interface	1-Regular	1-Regular	2-Micro	1-Micro

Table 1 Comparison of Controller Key Features

*It is important to note that a lot of the microcontrollers sold are copies of the real Arduino.cc boards. The schematics are readily available online. They can be cheaper than Arduino boards but the details can vary so you need to be a bit careful depending on your application. For instance, the Uno R3 pictured above is a copy. I've found them to generally function like the Arduino Uno R3 but the clock/crystal on the copy is 12 MHz versus 16 MHz on the Arduino.cc board. You'll need to take this into account if the code you are using is clock speed dependent.

As noted above, I used an Arduino Due for dynamic control of an ADF5355 evaluation board. The first time I compiled the code for this project I found out the Arduino Due doesn't have any EEPROM. It wasn't hard to add external EEPROM and it is a good example of the low-cost hardware add-ons available. I found small EEPROM boards with an I2C interface for less than \$2 each and adapted some open source code to interface it to the Due. Other examples of low-cost hardware you can get for your projects are shown in Figures 5 and 6. The LCD Display and Button Shield (Figure 5 2X16 LCD Display and Button Shield) plugs directly into the connectors on controllers like the Uno, Mega2560, and Due. This provides a handy human interface independent of a PC.

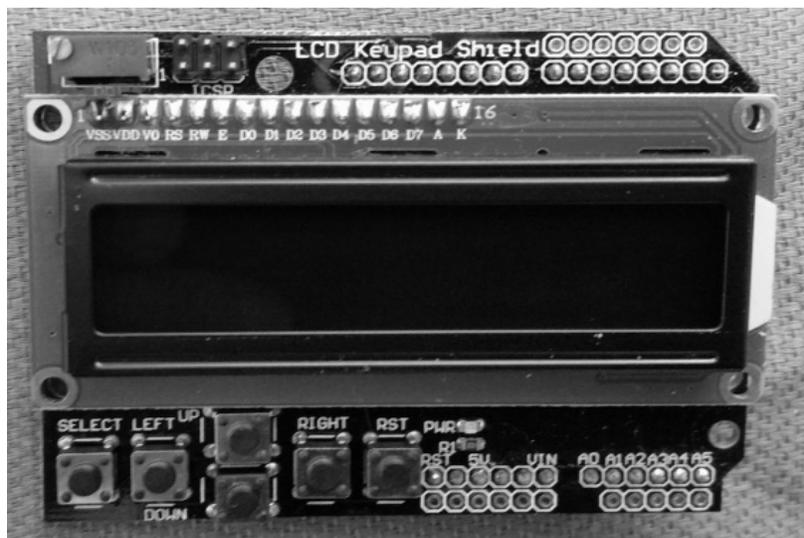


Figure 5 2X16 LCD Display and Button Shield

The relay boards shown in Figure 6 Arduino Compatible Relay Boards are handy for controlling devices not directly compatible with the controller I/O lines. The four-relay board on top is a shield and can plug directly onto the Mega2560, Uno, or Due, for instance. I did find a conflict on one of the I/O lines resulting in only three of the relays being useable when plugged into the controller for my application. The eight-relay board in the bottom of the figure is not a shield and although you don't have the convenience of a shield, you can assign whatever I/O pins you like to control the relays.

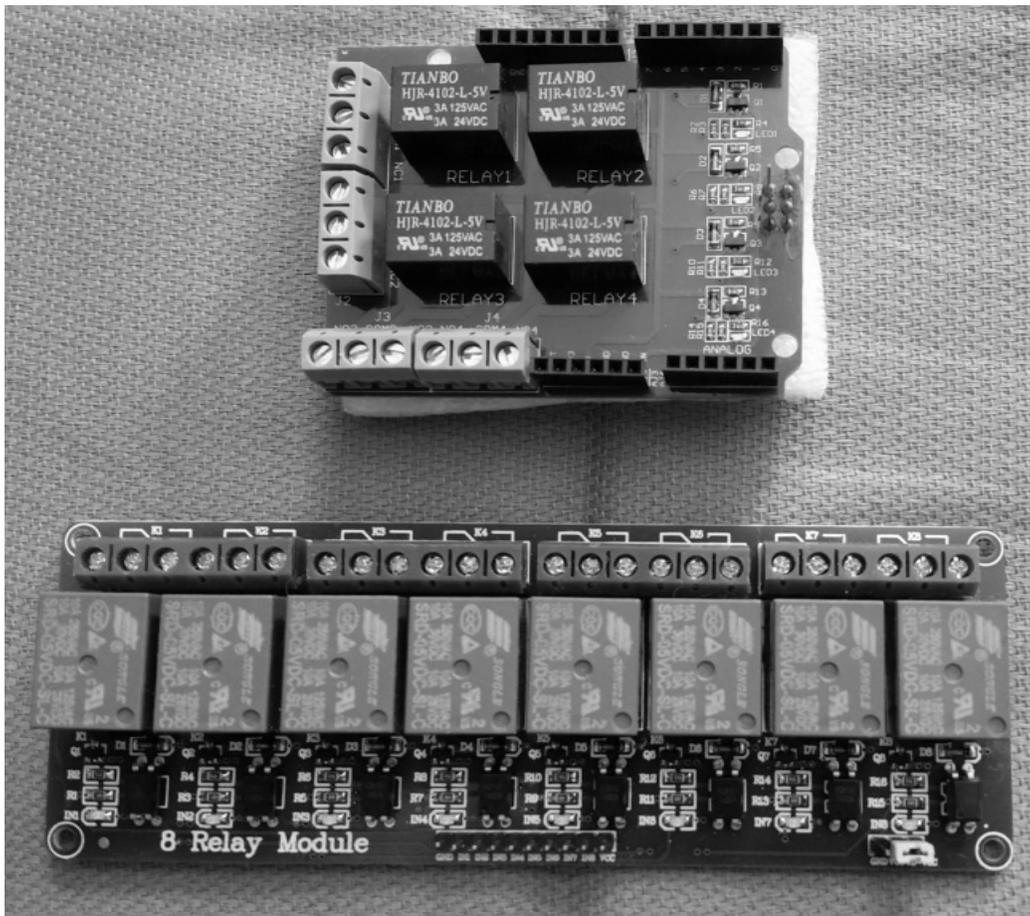


Figure 6 Arduino Compatible Relay Boards

There are many other devices that can be used with these controllers. I used the LM35 temperature sensor as part of the NTMS 432 MHz beacon project to send PA temperature as part of the beacon message. I've also experimented with LAN shields for local network use. An interesting note is that I found the software needed to go on the open internet (for virus protection, etc.) quickly gets too large for use with the smaller controllers. Prototyping shields, GPS modules, and other hardware too numerous to list are available for use with these controllers.

My final example is an in-work project using a 3-axis position sensor. The 3-axis sensor IC is a HMC5883L. I'm using the GY-271 version of the board and it connects to the controller via an I2C interface (Figure 7 3-axis Magnetic Sensor Interface to an Uno R3). The sensor is mounted to a small board and these can be found for just over \$2. The HMC5883L part is now obsolete, however, and some of the boards claiming to be based on the HMC5883L actually use copies of the part. It took a bit of detective work to get mine reading position since a different I2C I/O assignment was needed compared to the HMC5883L.

My goal is to have a more accurate readout of azimuth and elevation position for my 2M EME system. Anthony Good, K3NG with the help of others, has written Arduino code for a complete az/el control system. Information on the project and the code are available on the web (Reference 2). My particular challenge with this code is figuring out how to configure it since the code is flexible enough to be used with a variety of rotors, displays, position readouts, etc.

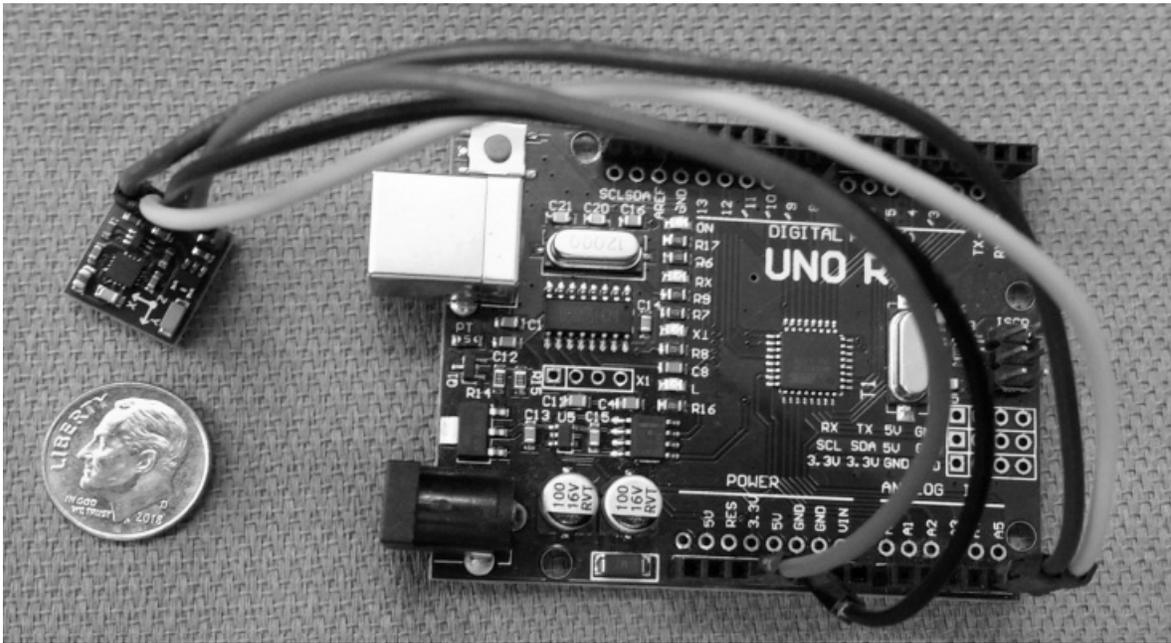


Figure 7 3-axis Magnetic Sensor Interface to an Uno R3

Current study is indicating that the GY-271 makes a good tilt-compensated electronic compass but an accelerometer will also be needed to get both azimuth and elevation readings. One candidate current part to do this is the TDK ICM-20948. It is a 9-degree of freedom (DOF) inertial measurement unit (IMU) that includes a:

- 3-axis gyro
- 3-axis accelerometer
- 3-axis compass
- And, a digital motion processor.

Another similar part is the STMicroelectronics LSM303D. Small boards with these and the supporting parts needed to interface to an Arduino are available for low cost from multiple sources. These parts use micro electromechanical systems (MEMS) technology. These or similar parts are used in smart phones and thus we have relatively cheap access to some powerful sensor technologies that can be exploited with Arduino or other microcontrollers.

These examples hopefully give you an idea of how Arduino or other microcontrollers can be used by the microwave amateur operator. The variety of controllers and add-on hardware available for low-cost is great. Also, the Arduino Integrated Development Environment (IDE) is available free and is a powerful tool for developing the code (sketches) and loading them onto the controllers (Reference 3).

References

1. "Arduino Controlled Microwave Frequency Sources", G. McIntire, AA5C, Microwave Update Proceedings 2017.
2. https://github.com/k3ng/k3ng_rotator_controller
3. <https://www.arduino.cc/en/Main/Software>

A decorative graphic on the left side of the slide, consisting of white lines and circles on a dark grey background, resembling a circuit board or a network diagram.

ENHANCED PROPAGATION MICROWAVE EVENTS

A PRIMER / WIDE AREA OPENING CASE-STUDIES

JOE JURECKA – N5PYK

NATIONAL WEATHER SERVICE

LUBBOCK, TEXAS

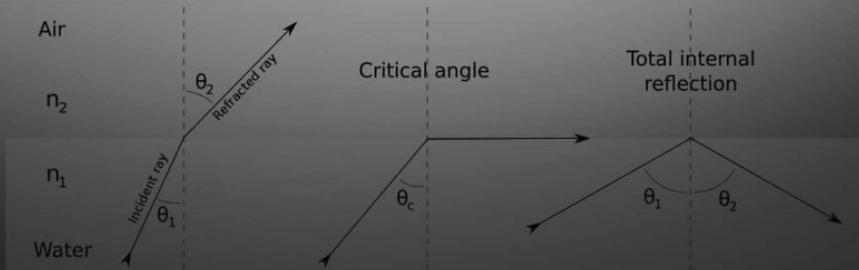
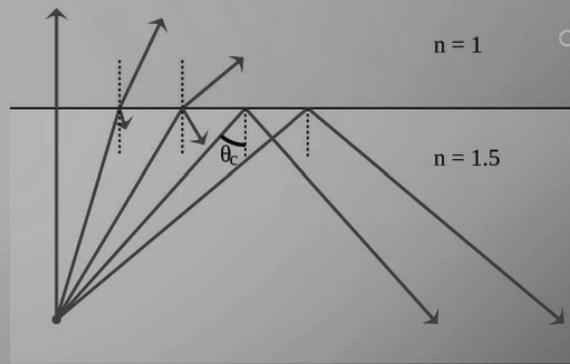
A decorative graphic on the left side of the slide, consisting of white lines and circles on a dark grey background, resembling a circuit board or a network diagram.

MOTIVATION

- Who doesn't like a great band opening?
 - How does it work?
 - What weather patterns support large-scale openings?
 - Why are some openings larger and stronger than others?
- 
- A decorative graphic on the right side of the slide, consisting of white lines and circles on a dark grey background, resembling a circuit board or a network diagram.

SNELL'S LAW

$$\frac{\sin\theta_2}{\sin\theta_1} = \frac{n_1}{n_2}$$



https://en.wikipedia.org/wiki/Snell%27s_law

K-FACTOR

Ratio of effective radius of radio wave ray vs. radius of earth (6370km)

$$K = \text{radius of propagation} / \text{radius of earth}$$

Radio horizon
Optical line of sight

Normally about a 1.4X distance multiplier on VHF.

Important for understanding enhanced propagation

SCHEMATIC OF PROFILE (REFRACTION) THE BASIS FOR OPENINGS

$$N = \frac{77.6}{T \left[P + \frac{4810(E_s)(RH)}{T} \right]}$$

‡FREEMAN

N=Refractivity

Es=Saturation Water vapor pressure (hPa)

P=Atmospheric pressure (hPa)

T=Temperature (Kelvin)

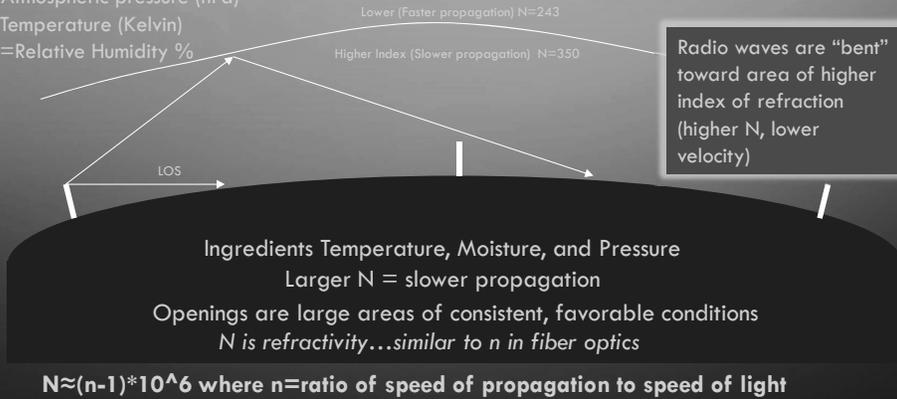
RH=Relative Humidity %

$$K \approx 1 / \{ [1 + (\Delta N / \Delta h) / 157] \}$$

K=K-factor

N=Refractivity

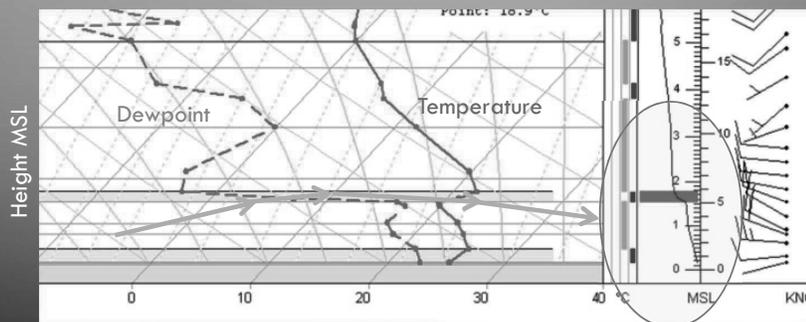
h=height (km)



5

LOCAL ENHANCEMENT (TROPOSPHERIC REFRACTION)

- Accomplished with single hop off inversion layer aloft
- Often a single thermal inversion layer aloft...sometimes very close to the ground
- Not generally spatially widespread



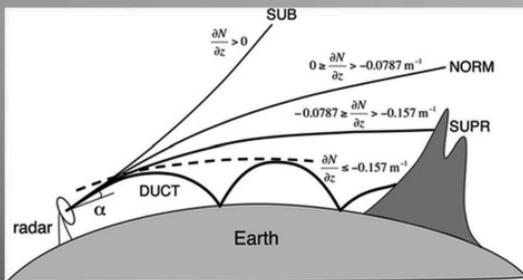
Very useful on microwave on spring and summer mornings and evenings!

6

REFRACTION MODES



REFRACTION CATEGORIES



The different propagation regimes of a ground-based radar beam emitted with a small tilt angle α above the horizontal plane: subrefraction (SUB), normal refraction (NORM), superrefraction (SUPR), and ducting (DUCT). The corresponding values of refractivity gradient $\partial N/\partial z$ are indicated above each beam path. The dashed line indicates the top of the duct.

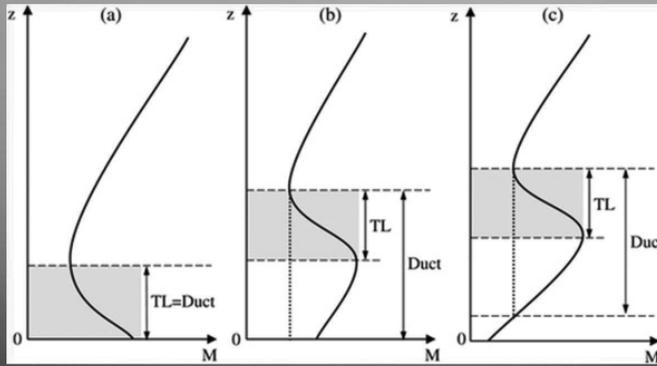
Change of N with height	Mode
$\partial N/\partial z > 0$	Subrefraction
$0 \geq \partial N/\partial z > -0.0787 \text{ m}^{-1}$	Normal refraction
$-0.0787 \geq \partial N/\partial z > -0.157 \text{ m}^{-1}$	Superrefraction
$-0.157 \geq \partial N/\partial z$	Ducting

Steiner and Smith (2002)

<https://journals.ametsoc.org/doi/full/10.1175/2008JAMC1961.1>

THERMAL PROFILES

Duct thickness is a factor for usable frequencies



. Duct types as a function of the shape of the M vertical profile: (a) surface duct, (b) S-shaped surface duct, and (c) elevated duct. Gray shading identifies the trapping layer (TL)

<https://journals.ametsoc.org/doi/full/10.1175/2008JAMC1961.1>
<https://journals.ametsoc.org/doi/pdf/10.1175/2010JAMC2415.1>

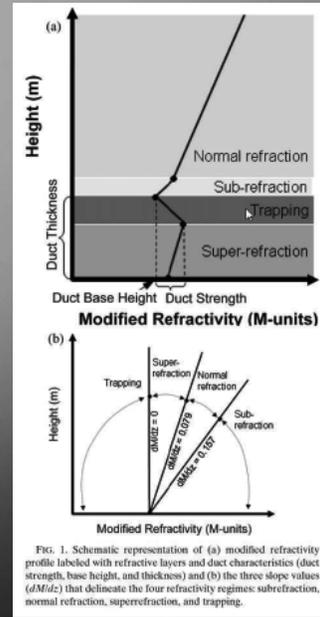


FIG. 1. Schematic representation of (a) modified refractivity profile labeled with refractive layers and duct characteristics (duct strength, base height, and thickness) and (b) the three slope values (dM/dz) that delineate the four refractivity regimes: subrefraction, normal refraction, superrefraction, and trapping.

INVERSION THICKNESS VS. FREQUENCY

Inversion Thickness		LUF	
Feet	Metres	Band	M/GHz
50	15	μ W	11.67 GHz
100	30	μ W	3.35 GHz
150	46	UHF	1615 MHz
200	61	UHF	962 MHz
300	91	UHF	464 MHz
400	122	VHF	276 MHz
500	152	VHF	185 MHz
600	183	VHF	133 MHz
700	213	VHF	101 MHz
800	244	VHF	79 MHz
900	274	VHF	64 MHz
1000	305	VHF	53 MHz
1100	335	VHF	45 MHz
1200	366	VHF	38 MHz
1300	396	VHF	33 MHz
1400	427	SW	29 MHz
1500	457	SW	26 MHz

From <http://www.dxinfocentre.com/propagation/luf.htm>

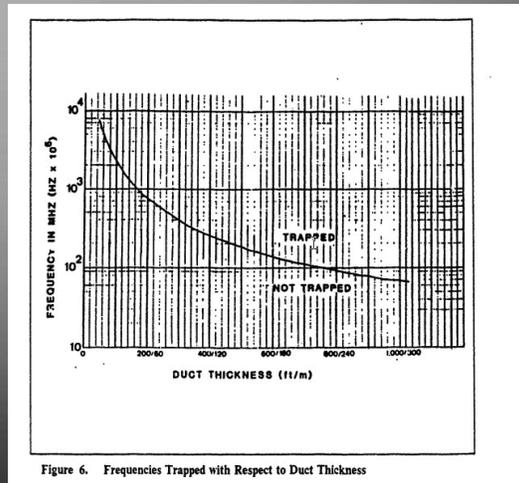
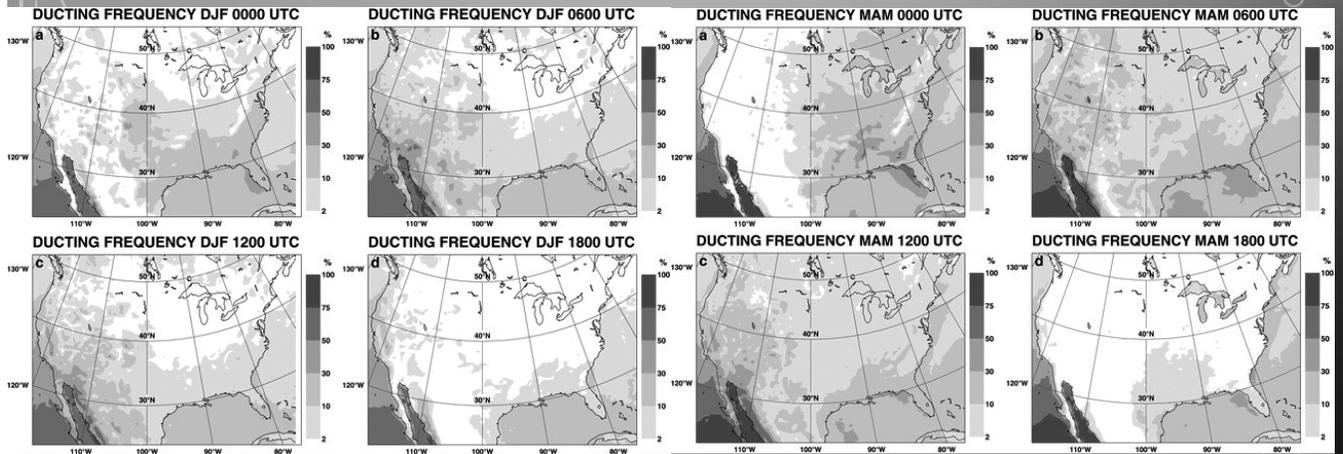


Figure 6. Frequencies Trapped with Respect to Duct Thickness

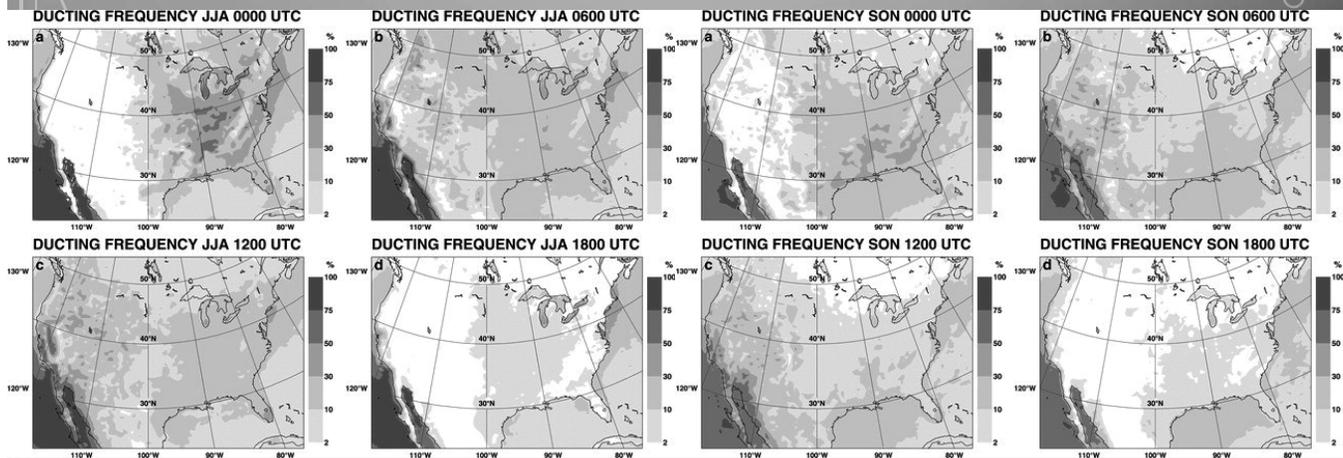
<https://apps.dtic-mil/dtic/tr/fulltext/u2/c034073.pdf>

DUCTING FREQUENCY



<https://journals.ametsoc.org/doi/full/10.1175/2008JAMC1961.1>

DUCTING FREQUENCY



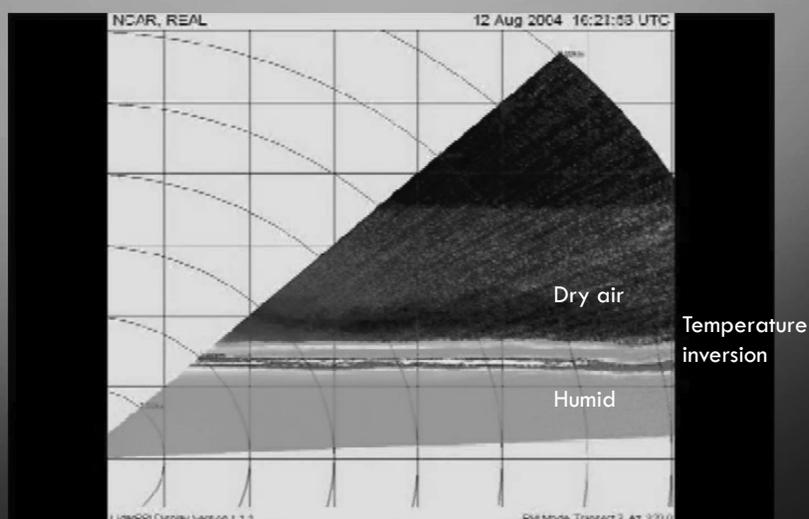
<https://journals.ametsoc.org/doi/full/10.1175/2008JAMC1961.1>

THE MIRAGE: A VISUAL ANALOGY FOR MICROWAVE REFRACTICION



• https://media.mnn.com/assets/images/2017/03/desert-mirage.jpg.638x0_q80_crop-smart.jpg

LIDAR VERTICAL CROSS SECTION

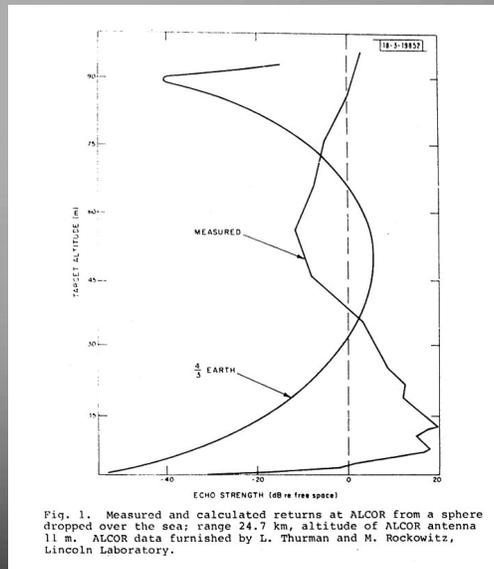


KWAJALEIN

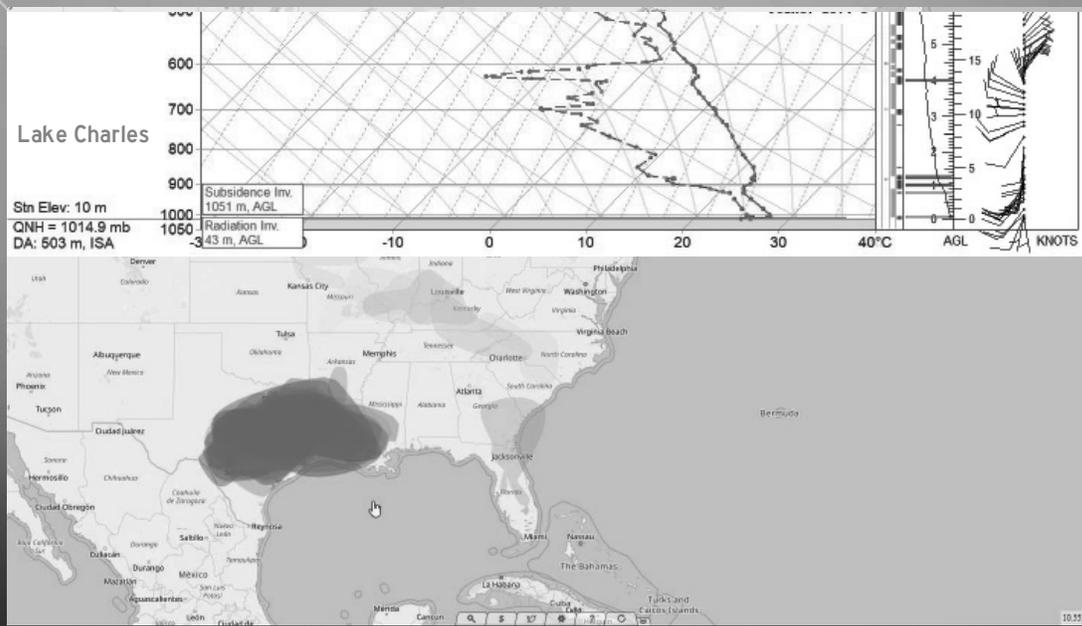
A sphere with a known radar cross section was dropped at a range of 25km

Note the sharp signal increase below 30m due to the low level evaporation duct over the sea

Was noted that frequencies below S band were not enhanced



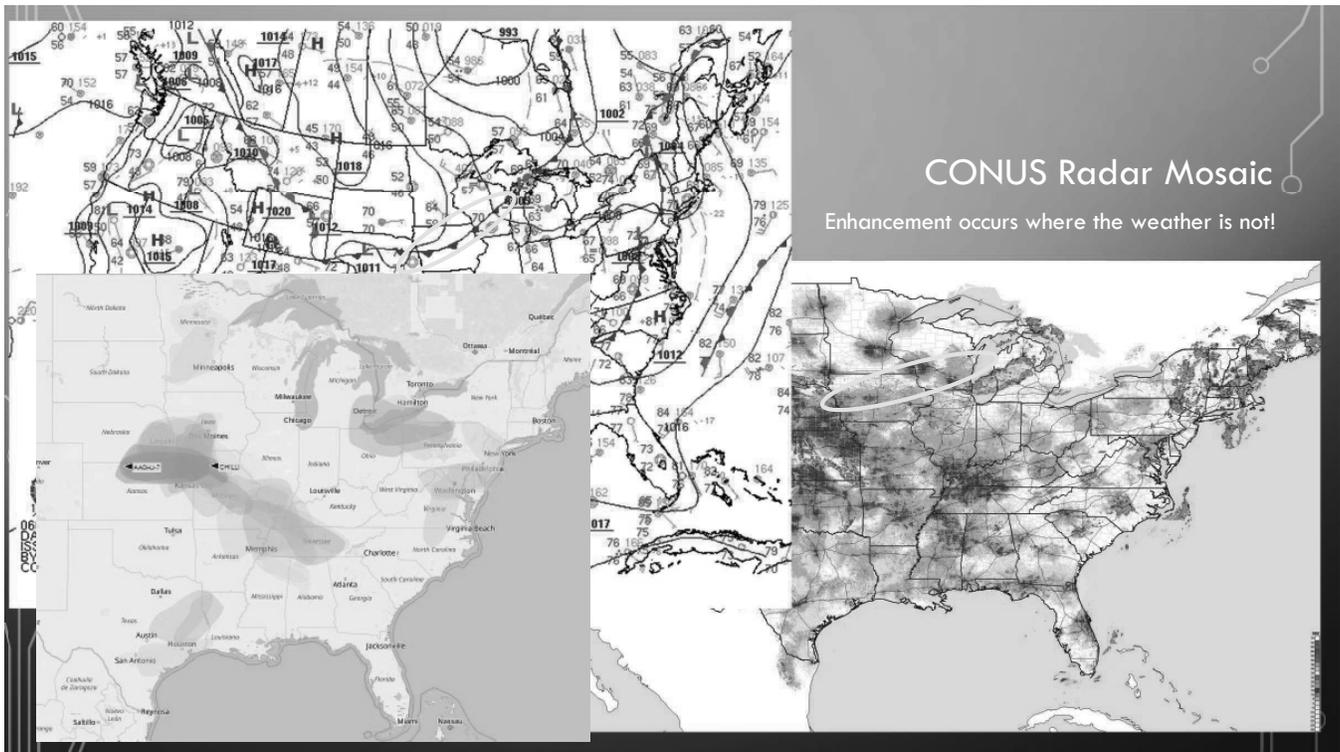
<https://apps.dtic.mil/dtic/tr/fulltext/u2/a057117.pdf>



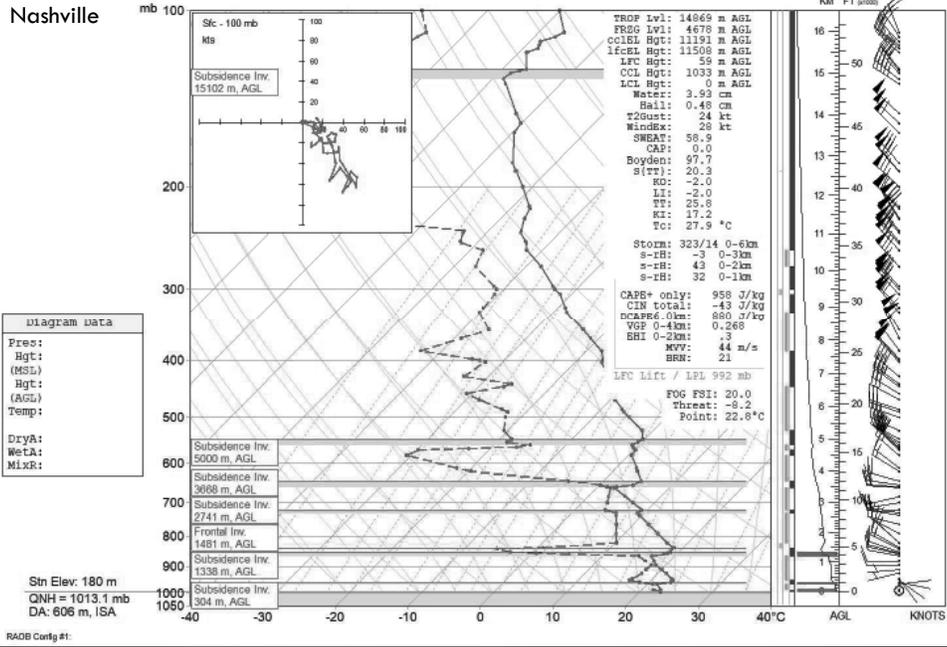
TROPOSPHERIC DUCTING

- Can enable contacts over distances of over a thousand miles.
- RF signal is propagated in duct
- Much variance exists on where signal exits the duct
- Signal can pass right over some users (elevated trapping ducts) such that it is possible to work stations from Texas to Florida but not into Mississippi

17

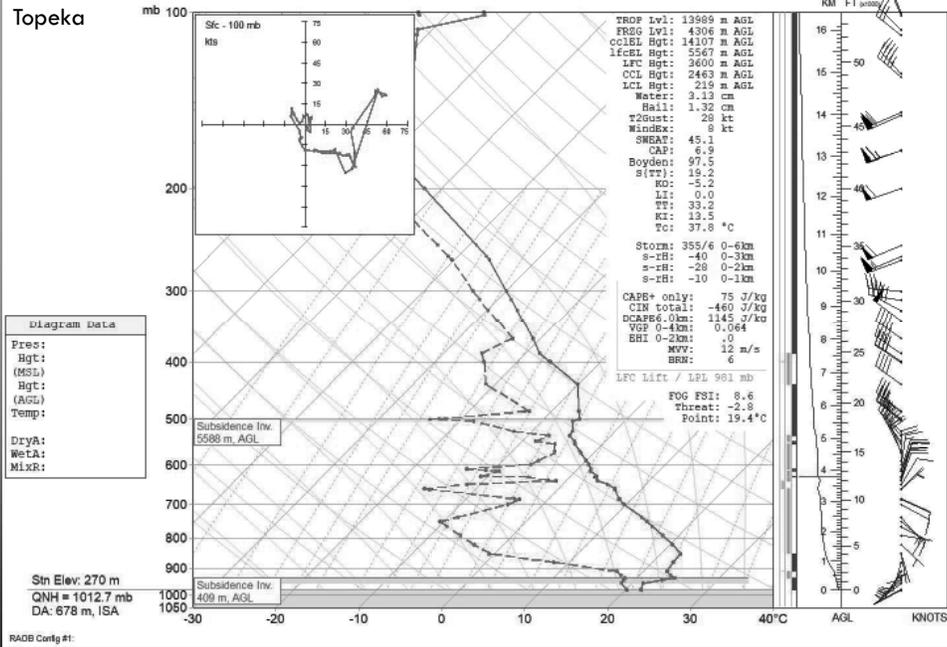


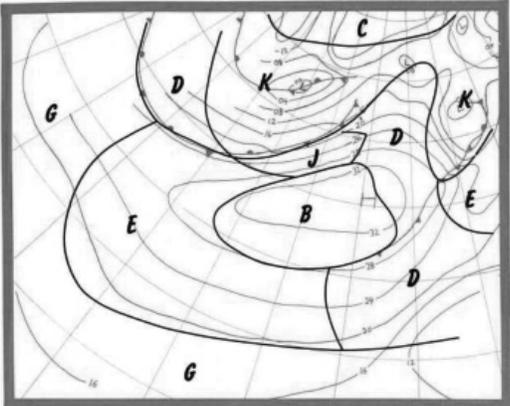
Nashville



<https://www.spc.noaa.gov/exper/soundings/>

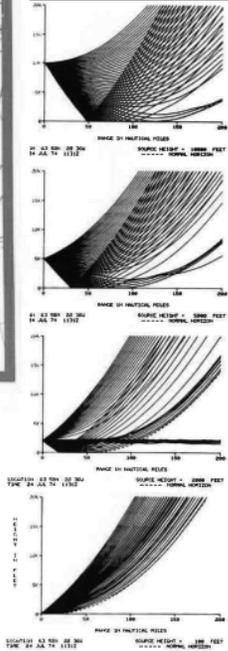
Topeka



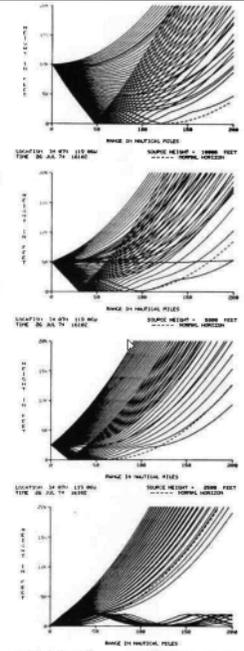
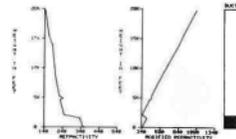


North Atlantic, Summer. High pressure covers most of the Atlantic; weak frontal systems traverse northern areas.

A-A Holes and fading possible for systems up to 10,000 feet. Ducting also possible for systems between 1,500 and 2,300 feet.



PROFILE TYPE B
Very strong surface-based duct with top near 2,000 feet. Moderate elevated temperature inversion. Subtropical air mass.



SYSTEMS COVERAGE
S-S Ducting and greatly extended range.
S-A Slightly extended range.
A-A Near normal for systems well above duct. Some holes and fading for systems slightly above duct. Ducting, holes and fading possible for systems within duct (below 2,000 feet).

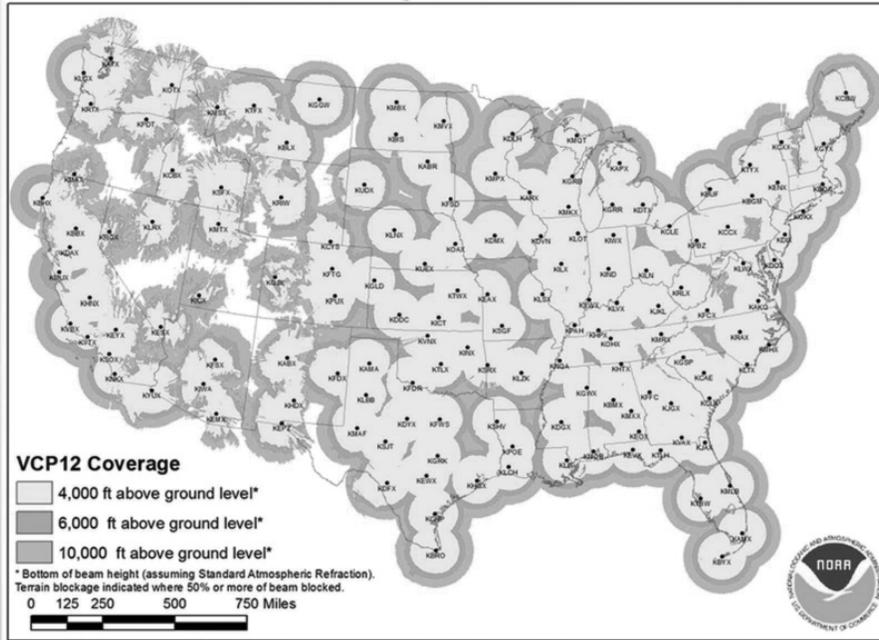
THE SOUNDING IS ONLY PART OF THE STORY

- Like in the mirage example, the weather must be very stable to propagate effectively.

Imagine the following analogy of an inversion (akin to a mirror)



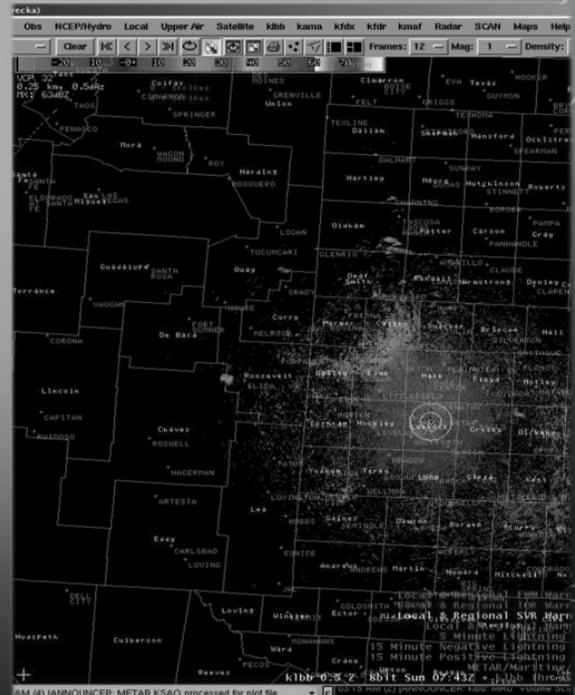
NEXRAD Coverage Below 10,000 Feet AGL



EFFECTS ON WEATHER RADAR

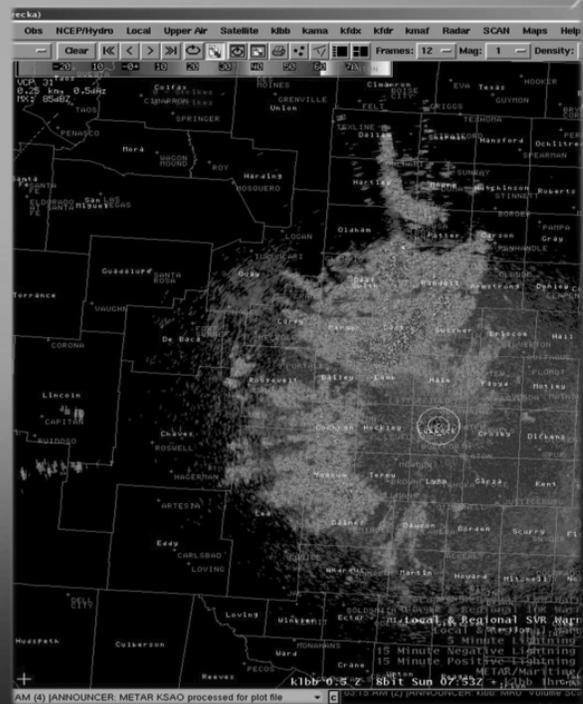
Image of WSR-88D radar data with Gaussian Model Adaptive Processing (GMAP) enabled under super-refraction conditions

$F_0 \sim 3$ GHz



EFFECTS ON WEATHER RADAR

10 minutes later with clutter suppression disabled



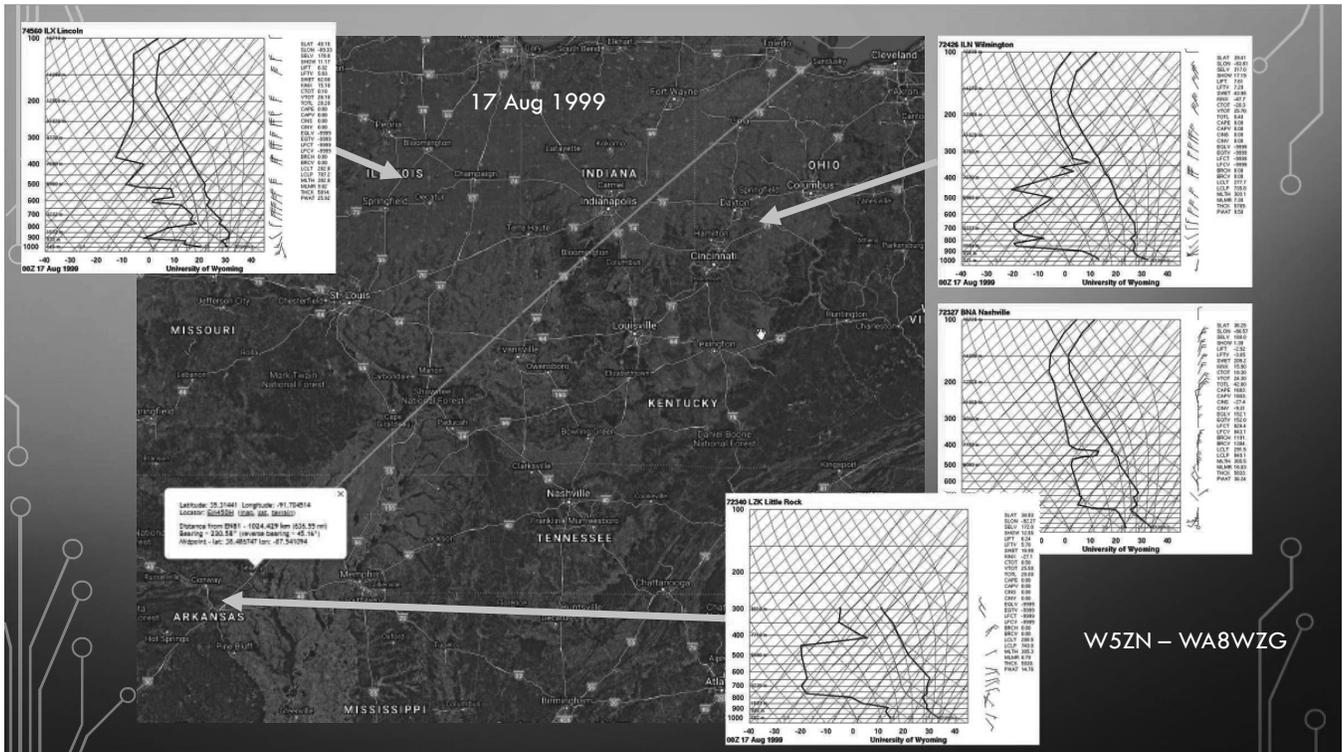
CASE STUDIES

10 GHZ AND UP

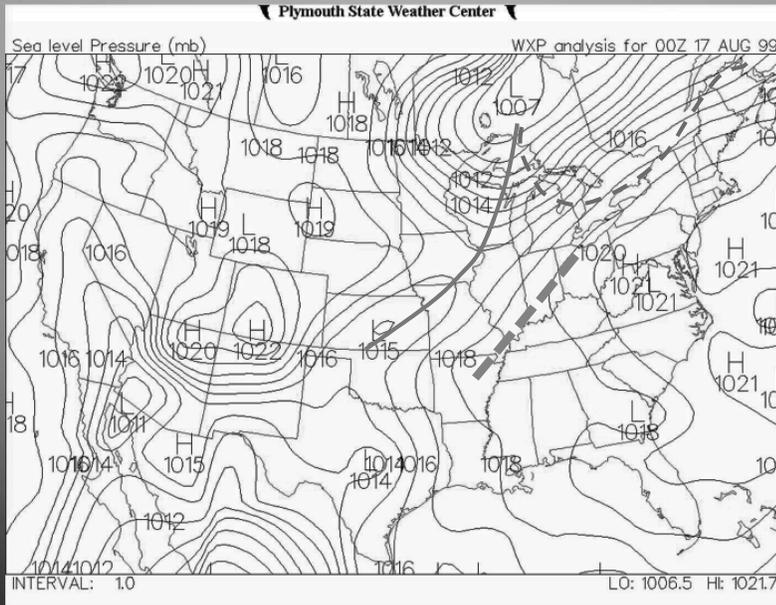
W5ZN (EM45dh)
to
WA8WZG (EN81om)

17 Aug 1999 (0032Z-0049Z)

DATE	FREQ	MODE	POWER	TIME	STATION WORKED	REPORT SENT	REC'D	TIME OFF	QTH	COMMENTS NAME	QSL VIA	QSL S	QSL R
8-15	144	USB	1kw	2303	W89ZE	59	59	2303	Illinois	Gary	EM45	✓	✓
8-16	"	"	"	1334	K8ZES	59	59	1334	New York	Jim	FA302	✓	✓
"	"	"	"	1334	W8BWIV	59	59	1335	Michigan	Tom	EN74	✓	✓
222	"	"	"	1339	W4MKE/9	59	59	1339	Indiana	Dave	EN7C	✓	✓
"	"	"	"	1340	VE2FU	59	59	1340	Canada	Steve	EN93	✓	✓
"	"	"	"	1341	K8ZES	59	59	1341	New York	Jim	FA302	✓	✓
144	"	"	"	1346	KF8DX	59	59	1346	Michigan	Tom	EN72	✓	✓
"	"	"	"	1352	N2CEI/2	59	59	1352	Ohio	Steve	EN81	✓	✓
432	"	"	"	1355	WA8WZG	59	59	1355	Ohio	Tom	EN81	✓	✓
1296	CW	150	"	1405	WA8WZG	599	599	1405	Ohio	Tom	EN81	✓	✓
2304	"	100	"	1411	WA8WZG	599	599	1411	Ohio	Tom	EN81	✓	✓
144	USB	1kw	"	1451	N2XTX	59	59	1451	New York	Dave	FA302	✓	✓
"	"	"	"	1455	VE2FU	59	59	1455	Canada	Steve	EN93	✓	✓
8-17	2456	CW	45	0032	WA8WZG	599	599	0032	Ohio	Tom	EN81	✓	✓
8760	"	"	"	0042	WA8WZG	599	599	0042	Ohio	Tom	EN81	✓	✓
"	USB	"	"	0043	WA8WZG	59	59	0043	Ohio	Tom	EN81	✓	✓
10368	CW	50	"	0049	WA8WZG	599	599	0049	Ohio	Tom	EN81	✓	✓
"	USB	"	"	0049	WA8WZG	59	59	0049	Ohio	Tom	EN81	✓	✓
432	CW	1kw	"	0100	W8RZF	559	599	0100	Ohio	Tom	EN91	✓	✓
"	USB	"	"	0106	K8CSD	59	59	0106	Ohio	Alex	EN81	✓	✓
144	"	"	"	0116	W8RZF	59	59	0116	Ohio	Alex	EN91	✓	✓
222	"	"	"	0118	W8RZF	59	59	0118	Ohio	Tom	EN91	✓	✓
"	"	"	"	0122	K8CSD	59	59	0122	Ohio	Alex	EN81	✓	✓
144	"	"	"	0123	K8CSD	59	59	0123	Ohio	Alex	EN81	✓	✓
"	"	"	"	0124	N8YQM	59	59	0124	Ohio	Doug	EN91	✓	✓

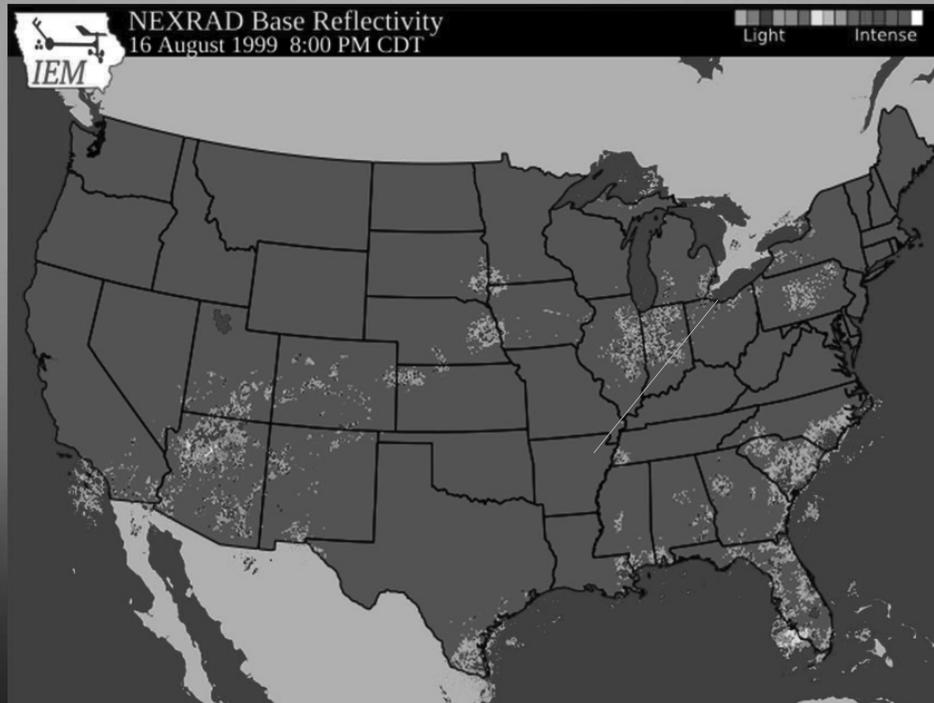


17 Aug 1999 0032Z



W5ZN - WA8WZG

<https://vortex.plymouth.edu/myo/sfc/ctrmap-a.html>



Surface Surface Dew Surface Visibility (mi) WXP analysis for 00Z 17 AUG 99

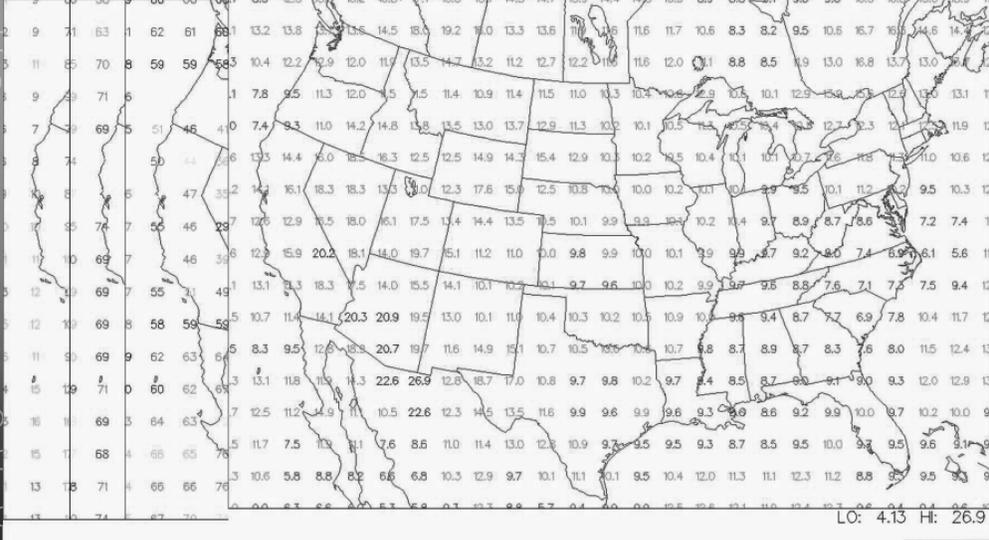
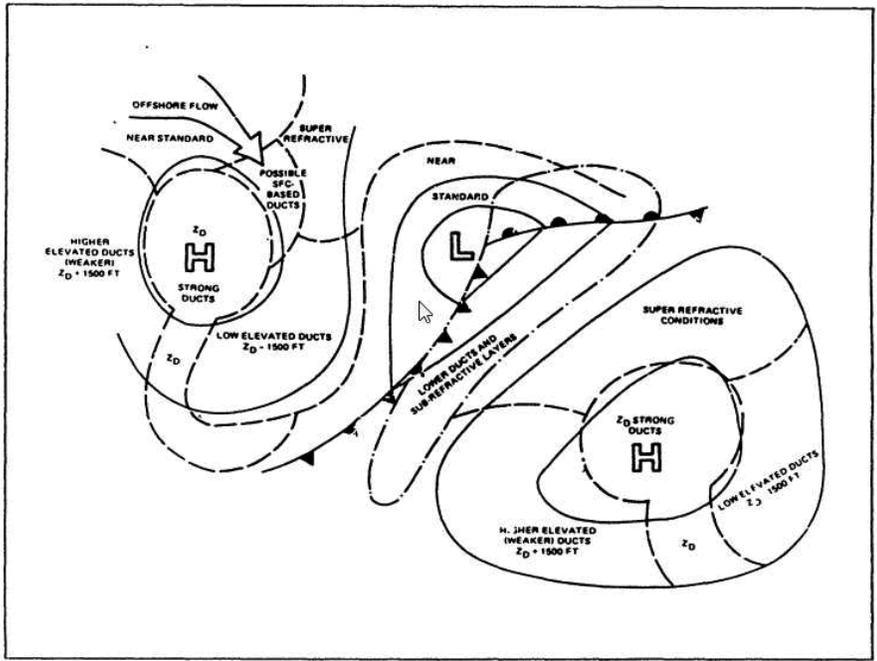
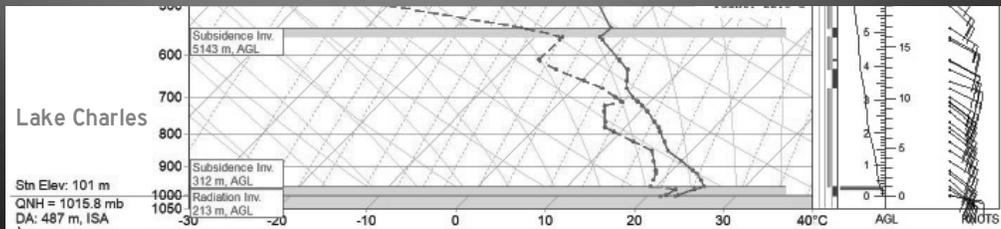
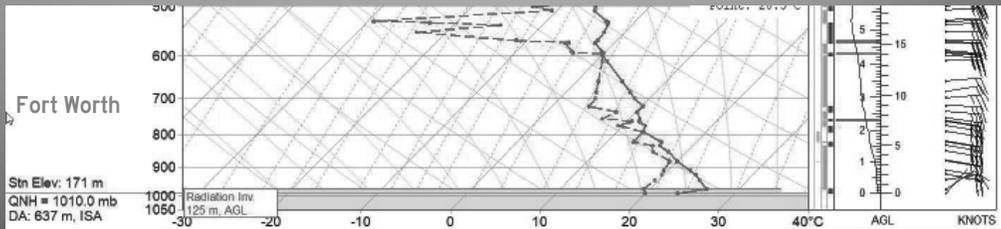
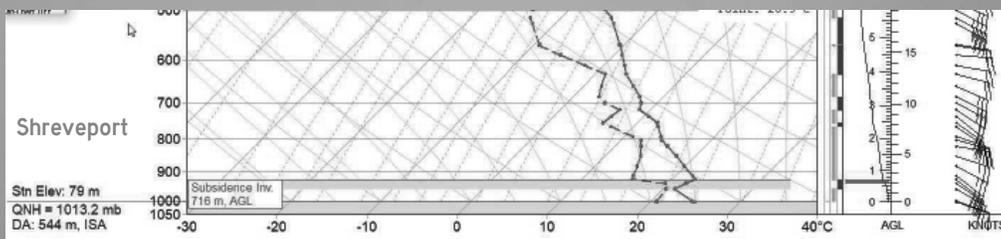
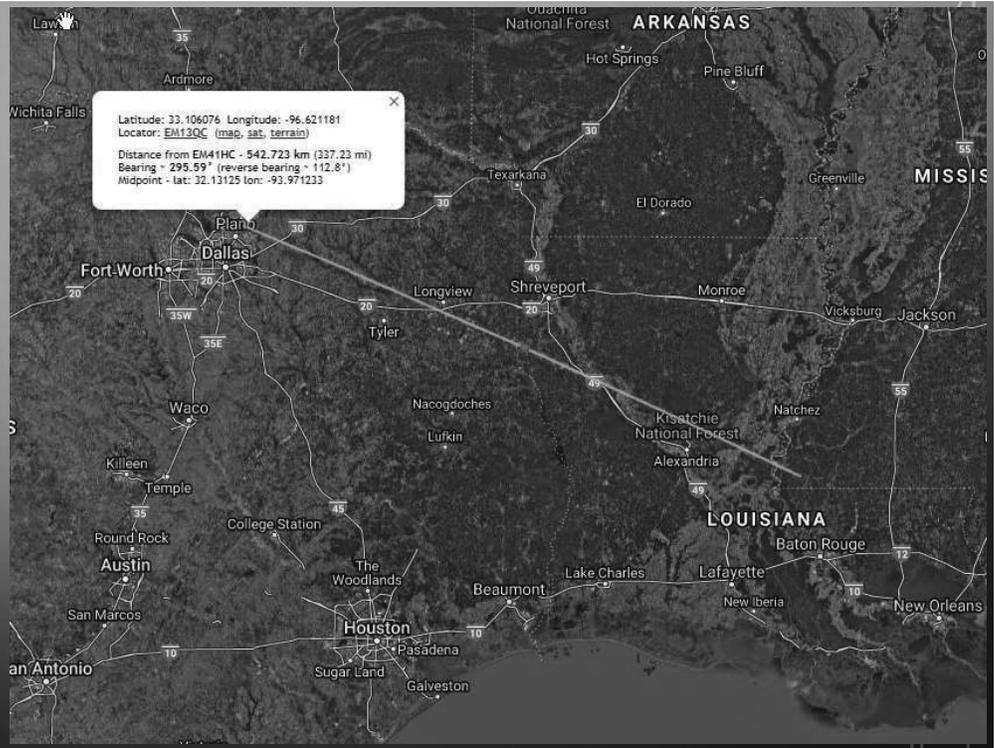


Figure 21. Model of Synoptic-Refractive Relationship



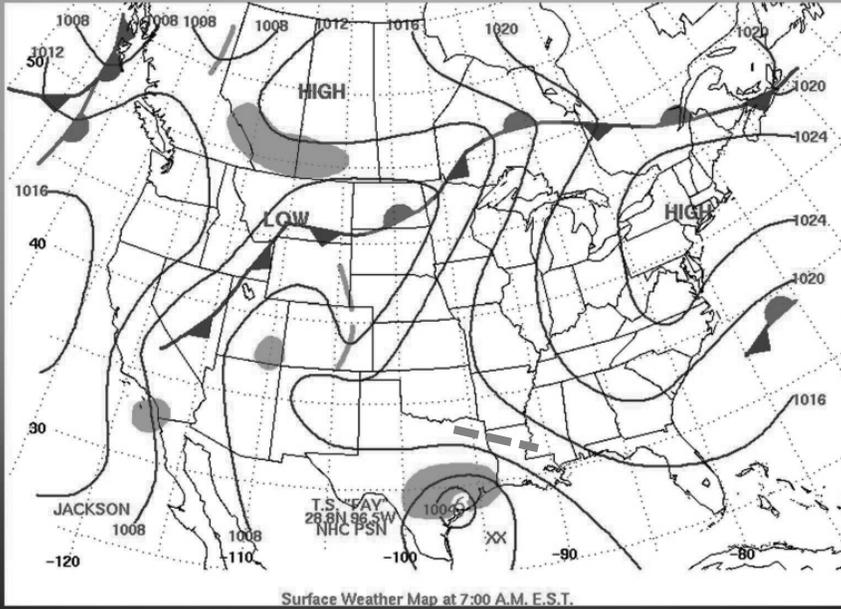
Culberson (1990)

7 Sept 2002
 24 GHz US Record
 W5LUA-WW2R
 1215Z
 10G & 24G



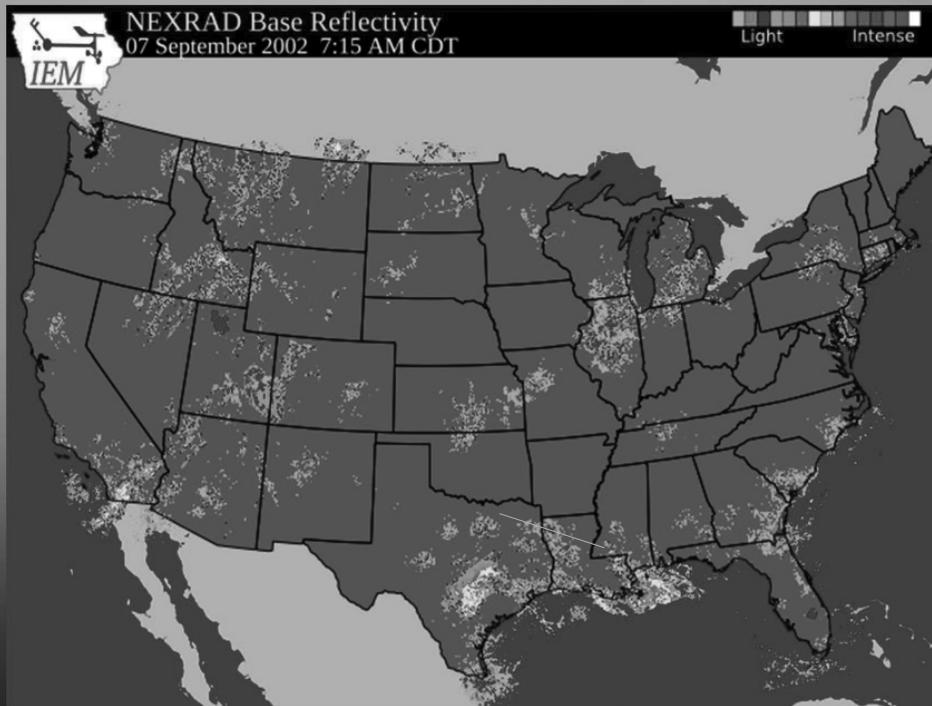
7 Sept 2002
 12Z

7 Sept 2002

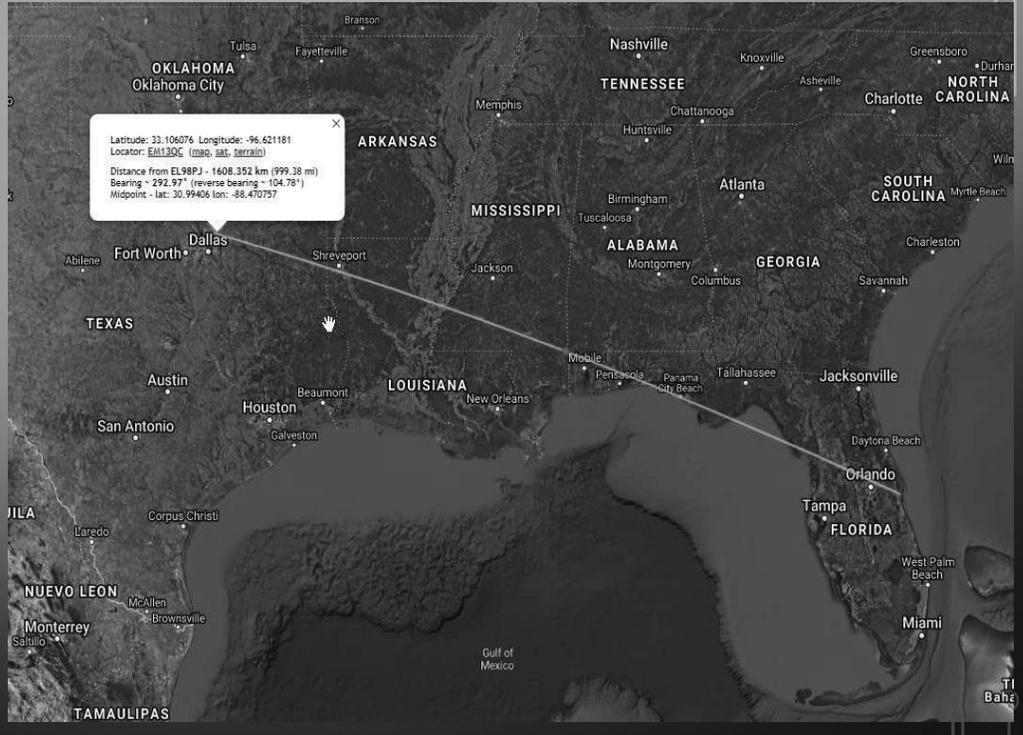


W5LUA - WW2R/R

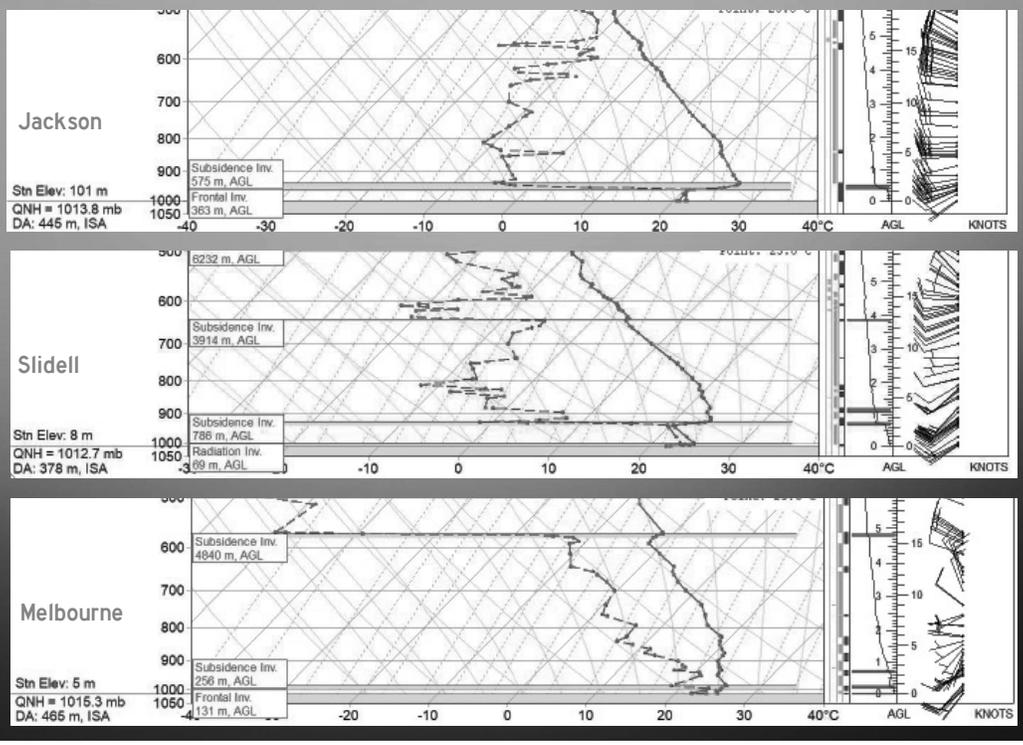
https://www.wpc.ncep.noaa.gov/dailywxmap/index_20020907.html

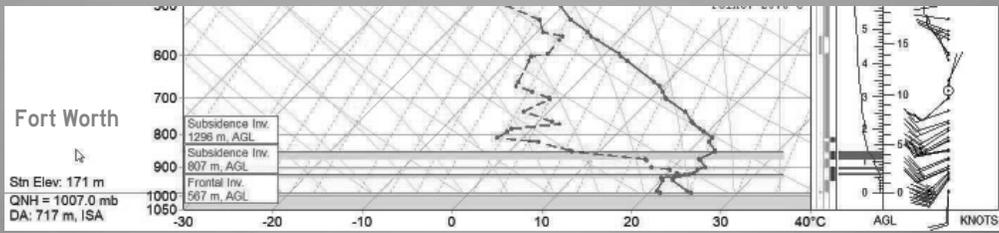


4 June 2012
 10 GHz
 W5LUA-KOVXM
 1148Z

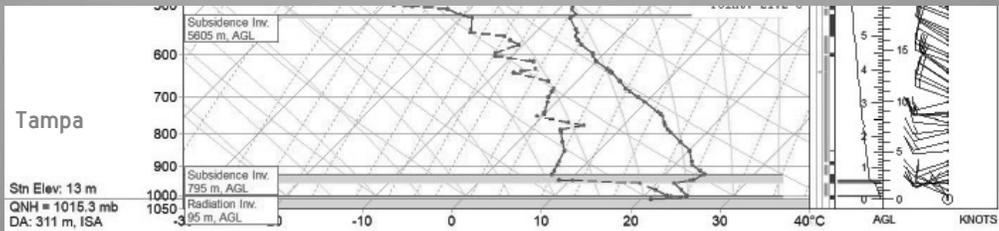
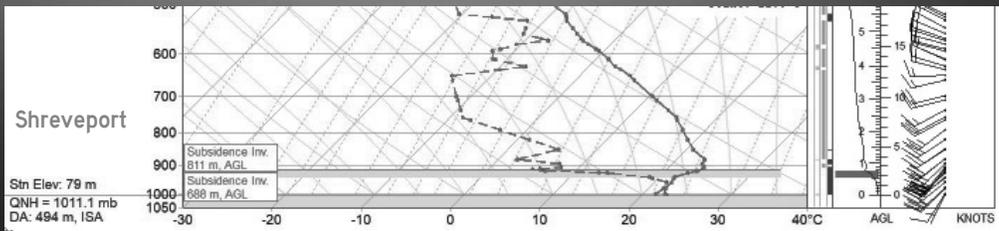
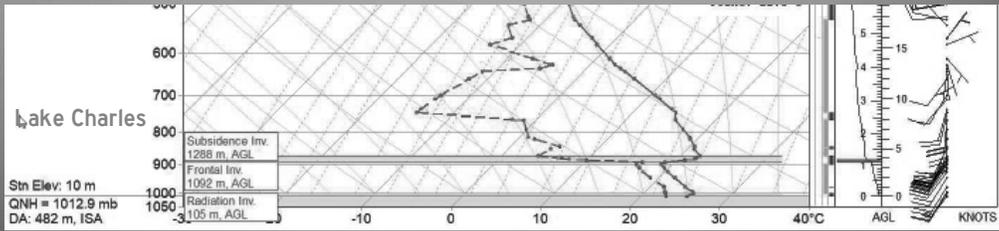


4 June 2012
 12Z

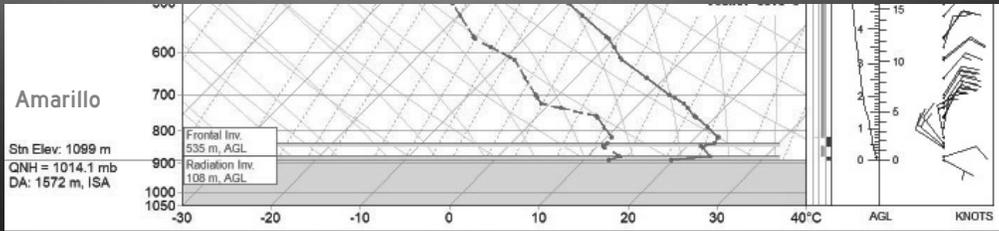
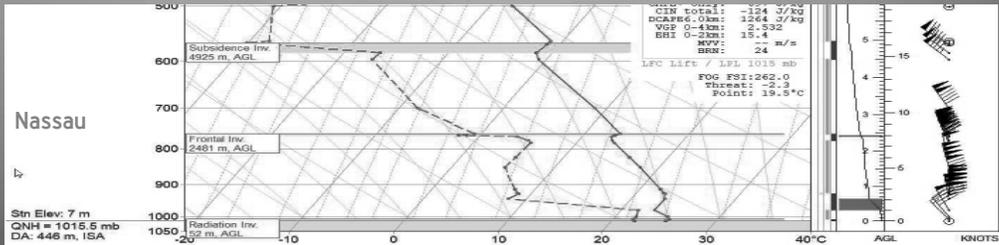




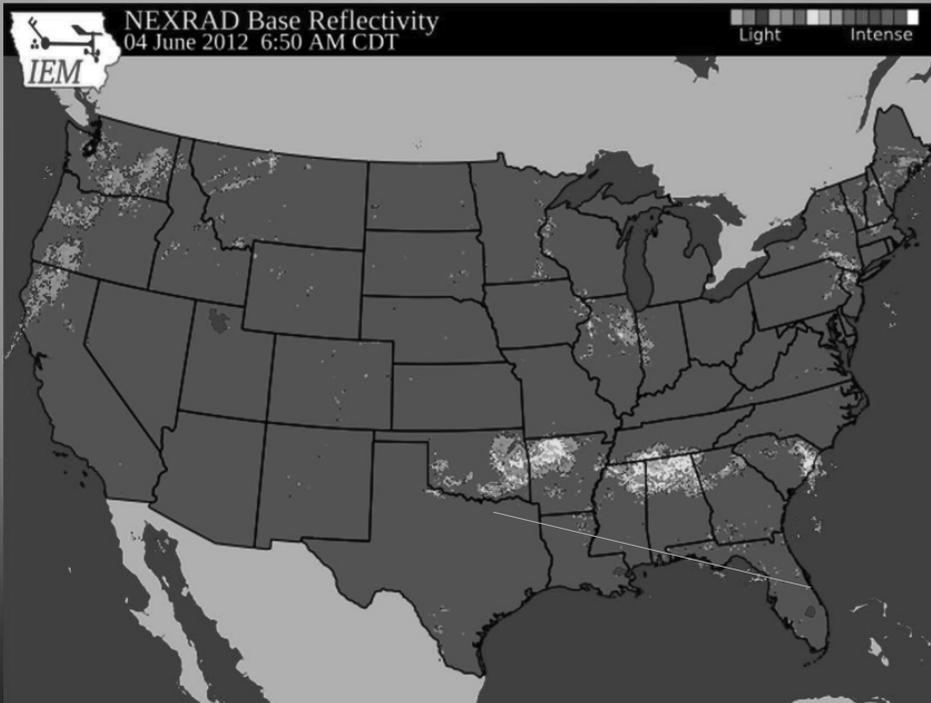
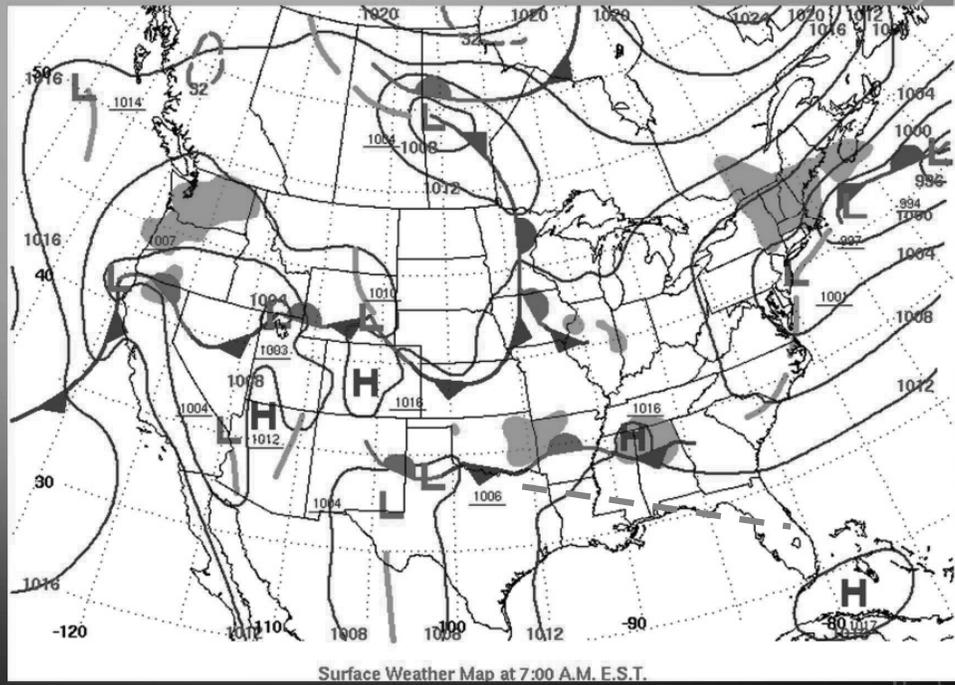
4 June 2012
12Z



4 June 2012
12Z



4 June 2012
10 GHz
W5LUA-KOVXM
1148Z



PACMISTESTCENT'S THUMB RULES (MARITIME) 1 / 3

Factors favorable to Elevated Duct Occurrence

1. Location within SE and SW quadrants of subtropical highs (for Bermuda area, SW and NW quadrants)
2. Anticyclonic curvature of surface isobars
3. Decreasing distance to center of high
4. Increasing surface pressure (especially PS > 1015 mB)
5. $T_{sfc} - T_{700} < 15^{\circ}\text{C}$ or $T_{700} = 5$ to 10°C
6. Location outside active frontal zone

These are maritime rules and some variations exist for over-land paths

Ref: TP 000005

PACMISTESTCENT'S THUMB RULES (MARITIME) 2 / 3

Factors favorable to Elevated Duct Occurrence

7. Presence of well-defined haze-layers
8. Presence of stratus clouds (not accompanied by rain. Drizzle from stratus is acceptable) *Note: Overland amateur contacts seem to favor clear skies*
9. Extensive stratus or stratocumulus sheet observed on visual or infrared satellite imagery with granular or cellular appearance
10. Evidence of a temperature inversion
11. Weak winds aloft
12. Lack of extensive and thick mid-level cloudiness

These are maritime rules and some variations exist for over-land paths

Ref: TP 000005

PACMISTESTCENT'S THUMB RULES (MARITIME) 3/3

Factors favorable to Surface Based (Non-evaporative) Ducts

1. Warm (temperatures higher than sea surface temperature), dry offshore flow
2. Stratus or fog deck with top at 1,000' or below
3. Large hole within stratus covered areas as observed on satellite imagery, or similar stratus-surrounded clear region extending seaward from continent
4. Stars or moon dimly visible through dense surface fog
5. Very smooth, white and uniform stratus observed on visual satellite imagery (as compared with more typical granular or cellular appearance)

Additional guidance and point system available within reference document

Ref: TP 000005

SUMMARY

Many factors must align for great microwave DX

- Vertical refractivity gradient
- Appropriate duct thickness
- Very stable duct region
- Homogenous conditions across a wide area
- Generally observed with mostly clear skies on-land

WHAT TO LOOK FOR LAND BASED OPENINGS

- Strong high pressure areas generally to the east of path
- Draping Stationary front north of the desired path
- Low surface wind speeds (often < 10 mph)
- Moist surface conditions (often dewpoints > 59°F (15°C) or higher)
- Temperature inversion in soundings
- Evidence of ducting in thermal profile

SPECIAL THANKS TO...

W5LUA - Al Ward

W5ZN – Joel Harrison

REFERENCES

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- Lopez, P., 2009: A 5-yr 40-km-Resolution Global Climatology of Superrefraction for Ground-Based Weather Radars. *J. Appl. Meteor. Climatol.*, **48**, 89–110, <https://doi.org/10.1175/2008JAMC1961.1>
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- Naval Electronics Laboratory Center Technical Note 3263, *An Evaluation of the Refractive Effects Guidebook (REG)*, by Glevy, D.F. and Logue, L.E., 02 December 1976.
- Commander third Fleet TACMEMO 280-1-76, *Refractive Effects Guidebook*, by J.S. Rosenthal, .. September 1976.
- Pacific Missile Test Center Technical Publication TP 000005, *Guide for Inferring Refractive Conditions from Synoptic Parameters*, by Helvey, R.A. and Rosenthal, J.S., 14 March 1983.

Practical applications of the Lime SDR Mini

Dave Robinson G4FRE WW2R
Lewisville 2019

Topics to be covered:-

Wideband SDR Transceivers
Microwave Narrowband
Eheilsat Narrowband
Portsdown
Eheilsat Wideband
Microwave DATV

Wideband SDR Transceivers

Past experience with SDR radios

1. Softrock
2. OPEN HPSDR Mercury/Penelope/Atlas Bus
3. Anan-10

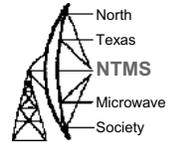
All above are HF radios

Requirements:-

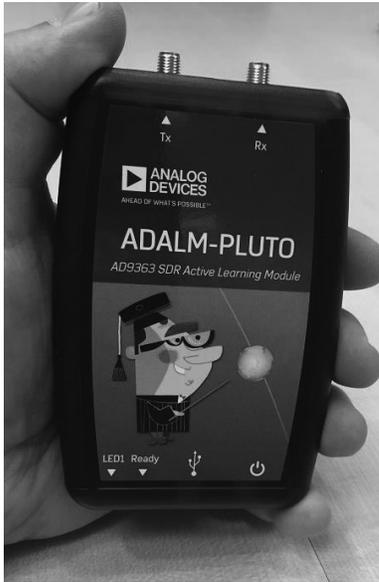
1. Transmit capable
2. Coverage of microwave bands
3. Software available

Shortlist:-

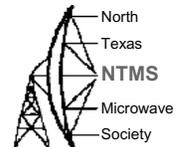
LimeSDR USB
LimeSDR Mini
Adalm Pluto
HackRF



Adalm Pluto



325 to 3800 MHz
 61.44 MSPS
 20 MHz bandwidth
 0.5mW out on a good Day!
 USB2 connection

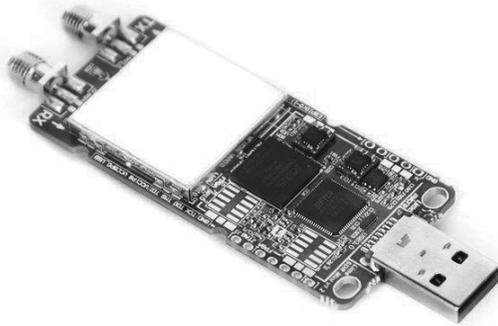
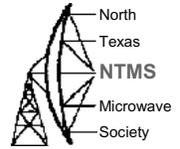


HackRF



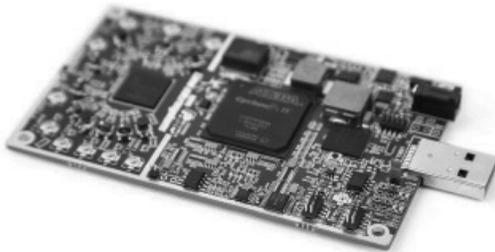
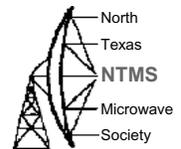
1MHz to 6 GHz
 20 MSPS
 20 MHz bandwidth
 1mW out
 8 bit
 USB connection
 \$300

LimeSDR Mini



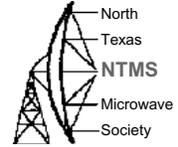
10 to 3500 MHz
61.44 MSPS
30.7 MHz bandwidth
1mW output
12 Bit
2 antenna ports
USB3 connection
\$160

LimeSDR



100k to 3500 MHz
61.44 MSPS
61.44 MHz bandwidth
~1mW output
12 Bit
8 antenna ports
USB3 connection
\$300

BEWARE!



When changing frequencies/power levels the SDR does a calibration routine. This can result in large spurious signals in and out of band!

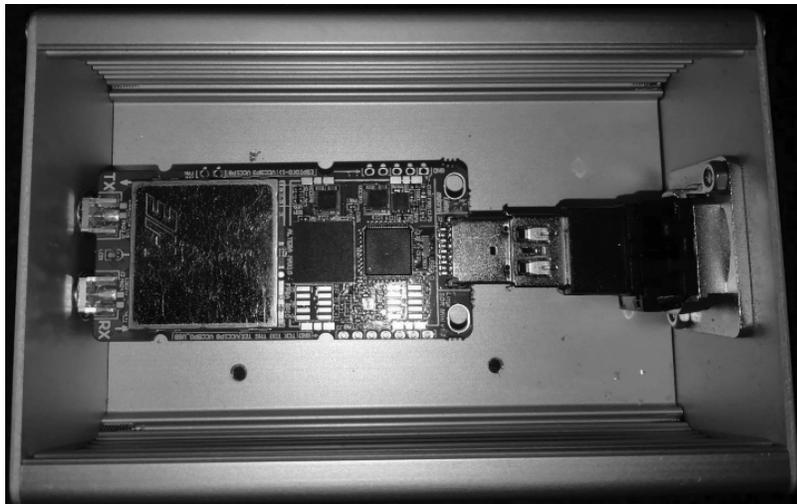
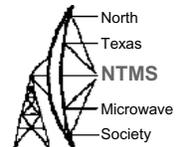
Example: Transmit carrier on 1296MHz

Adalm pluto generated a 10dbm signal randomly within its operating range

Lime SDR mini generated a 10dbm signal around 1296MHz before settling on desired power

Switch antennas and amplifiers appropriately!

I chose the LimeSDR Mini



USB Connector Very Fragile. Needs to be boxed

Microwave Narrowband

Microwave Narrowband Considerations

Can be used as a Beginners Transceiver covering 902, 1296, 2304 and 3456 MHz bands

Need to add Preamps / Filters / Amplifiers along with associated RF Switching

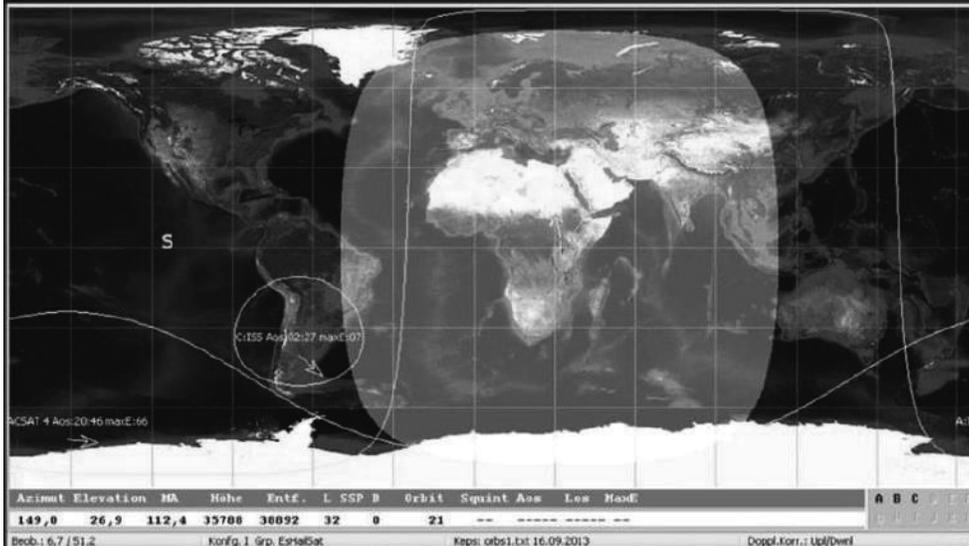
Use SDR console by G4ELI or SDR Angel by F4EXB Software

Phase noise and strong signal handling can be an issue in high activity areas

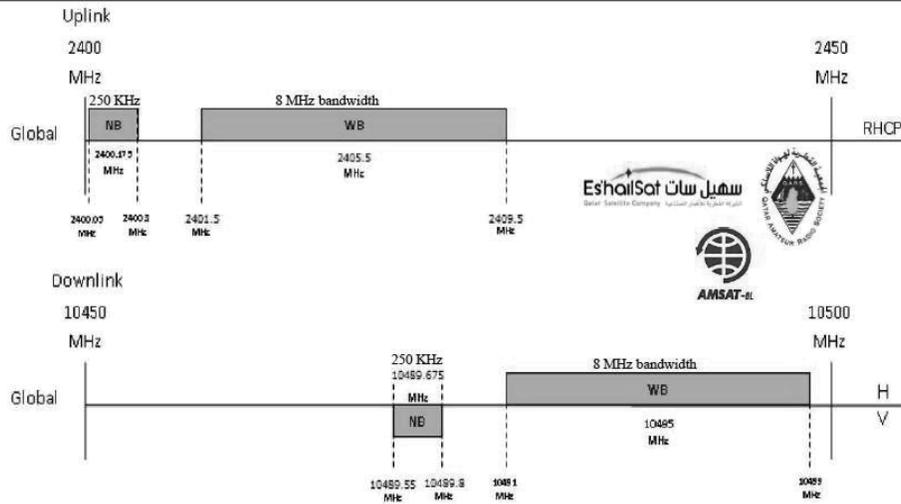
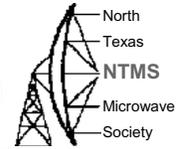
My barefoot pair of Limisdr mini have worked 33km on 13cm!

Es'HailSat-2

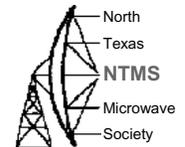
Earth Coverage Es'HailSat-2



Es'HailSat-2 Transponder Frequency Plan

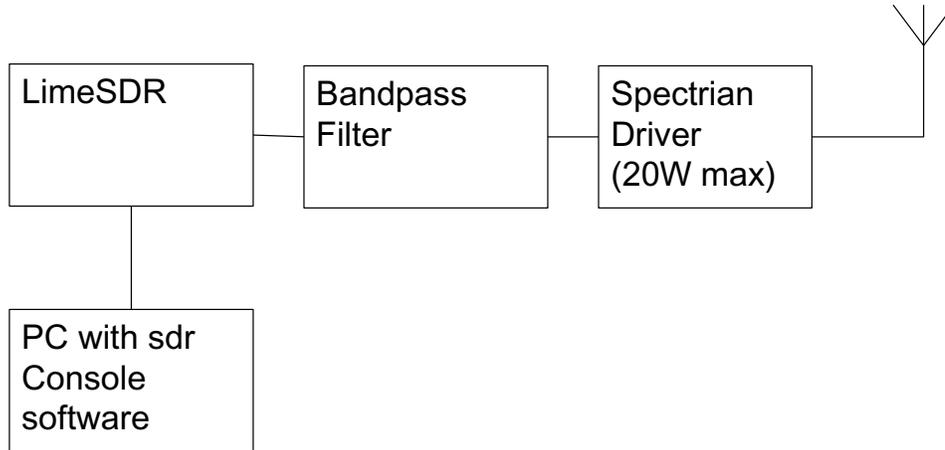


Xpdr	U/L FREQUENCY (MHz)			D/L FREQUENCY (MHz)			LO (MHz)	BW (MHz)		
	No	Pol	Center	Pol	Center	Center				
NB	RHCP	2400.05	2400.175	2400.3	V	10489.55	10489.675	10489.8	8089.5	0.25
WB	RHCP	2401.5	2405.5	2409.5	H	10491	10495	10499	8089.5	8



Es'HailSat-2 Narrowband

Es'HailSat-2 Narrowband Transmit

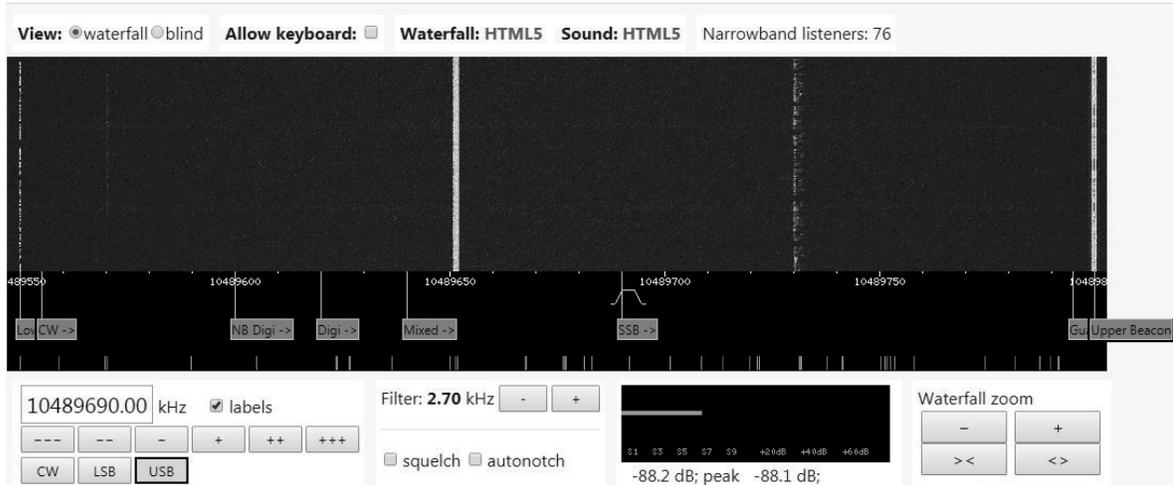
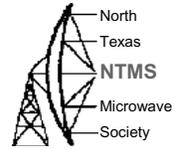


Uplink feed



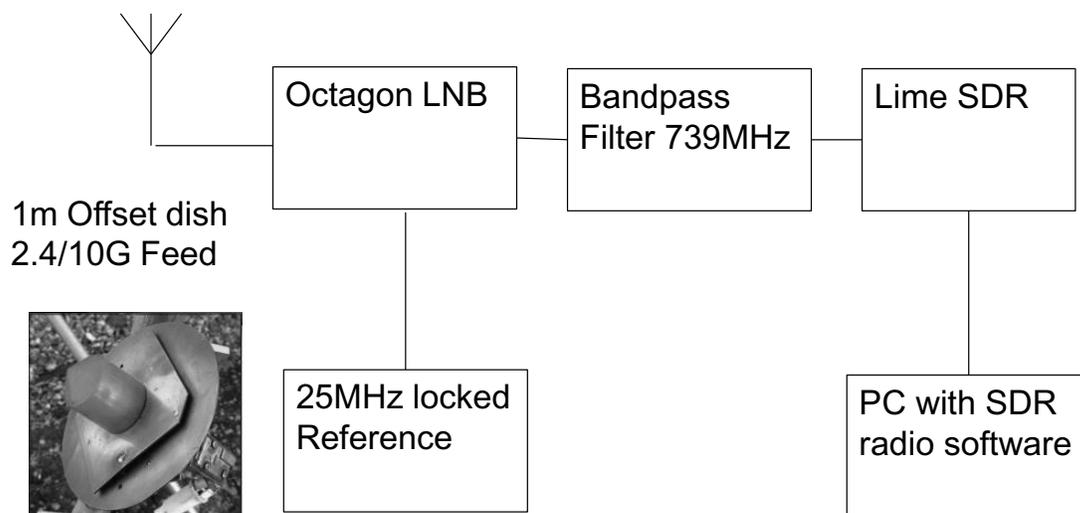
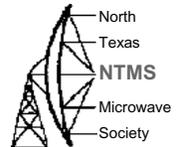
Needed on 4w with antenna pointing out of window!

BATC Narrowband WebSDR

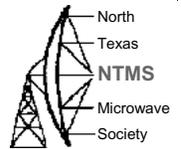
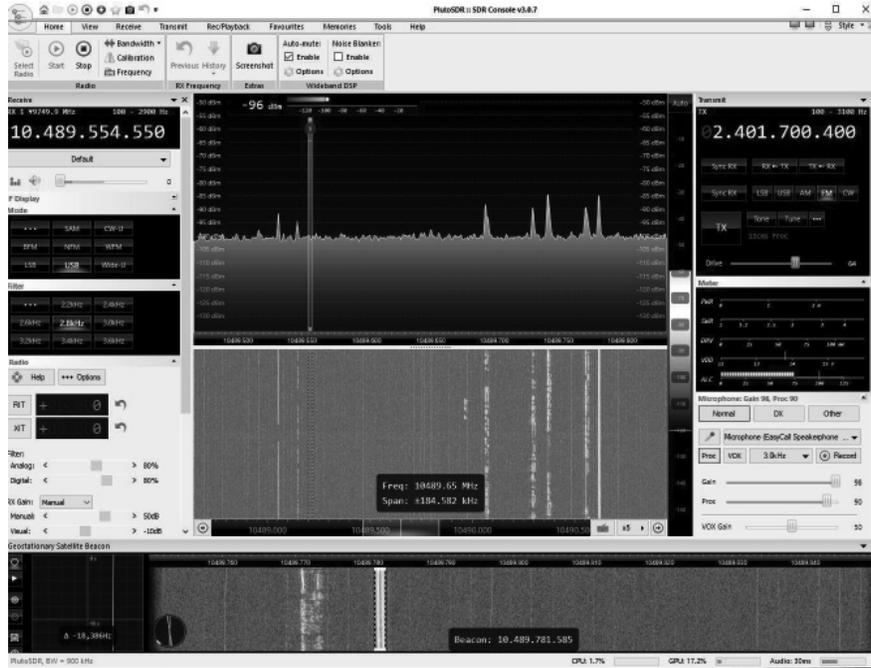
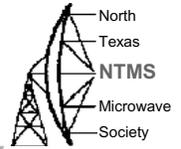


<https://eshail.batc.org.uk/nb/>

Downlink Receive

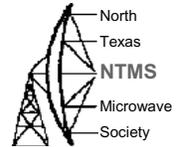


SDR Console Software



Portsdown DATV Exciter

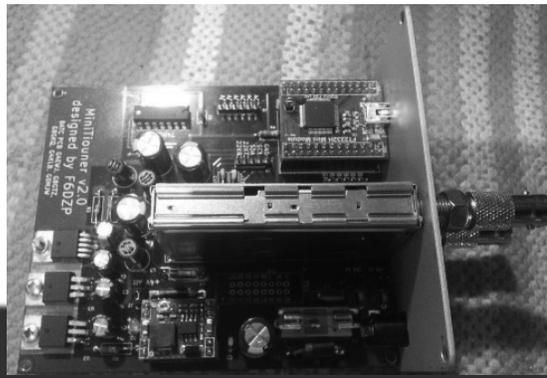
DATV Receiver



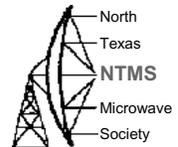
British Amateur Television Club (BATC) trying to promote Digital ATV (DATV)

Receiver design using wideband (146MHz to 2.4GHz) SATTV tuner integrated into Minituner Receiver hardware

Minituone software on PC by F5 takes receiver data via USB and displays DATV picture



Portsdown DATV Exciter



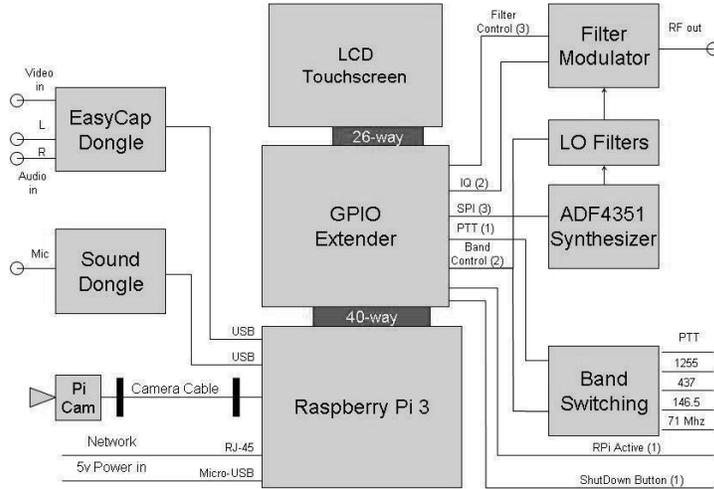
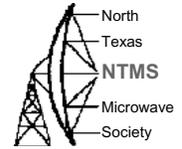
Transmitting not so Easy!

Some Commercial products existed but were considered too expensive for the masses and/or no longer in Production

BATC came up with their own design ("Portsdown") based on a Raspberry Pi 3

Software by G8GKQ

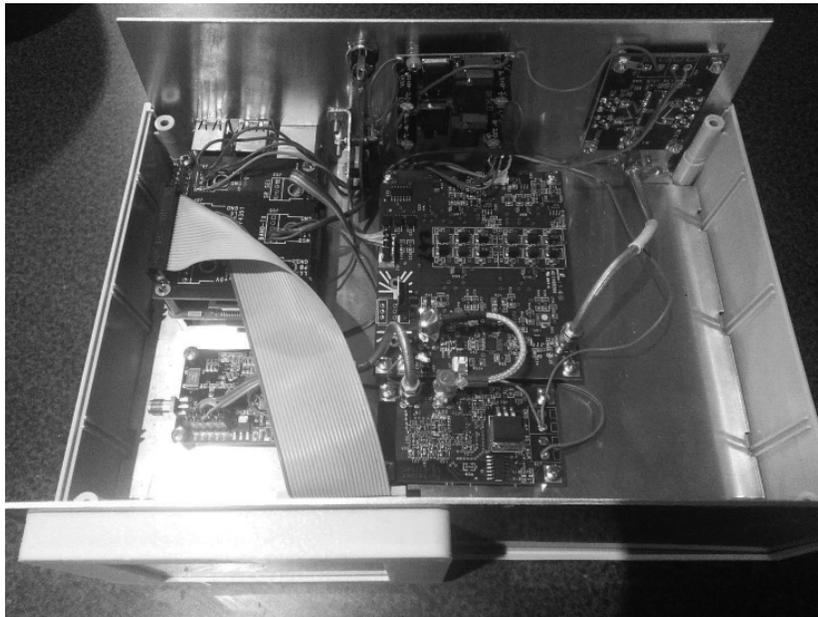
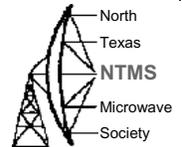
Original Portsdown 2018



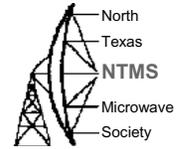
The Portsdown ATV Transmitter

080KQ 4 Feb 17

Original Portsdown 2018



Portsdown Next Generation



Critical Part of design was the Frequency modulator.
 Built and aligned board was made available from BATC

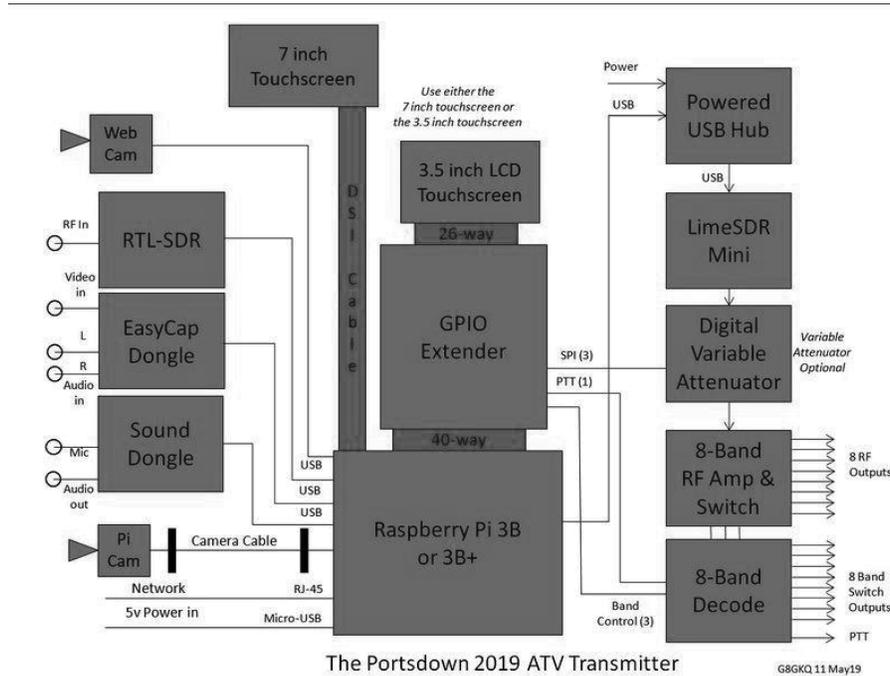
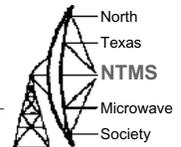
Parts sold to 350+ builders including USA+

Design needed updating especially as insufficient demand for another batch of 100FM boards

It was decided that next generation would be SDR based

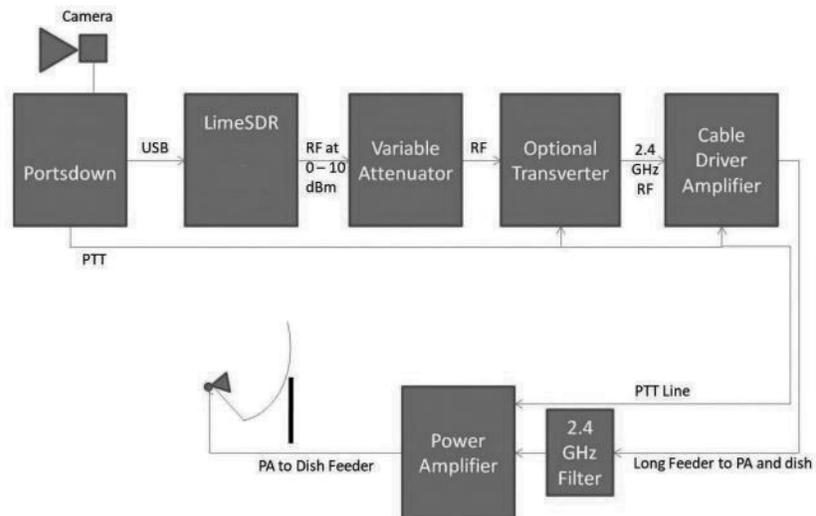
LimeSDR was chosen as software design was easier as better supported

Portsdown 2019

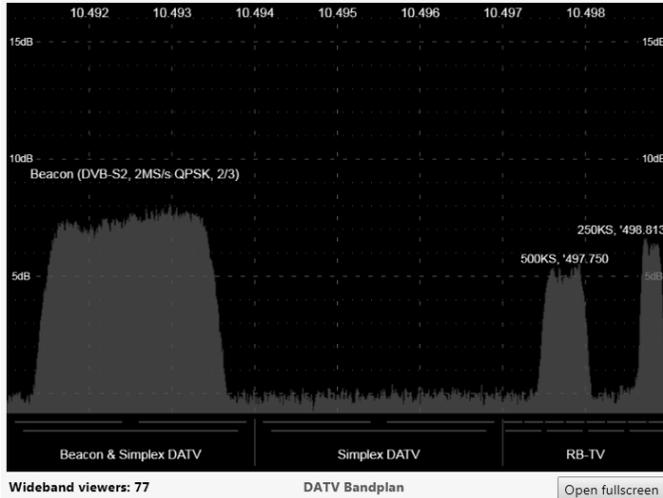
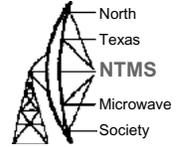


Es'HailSat-2 DATV

DATV Uplink Transmitter



BATC Wideband Spectrum viewer



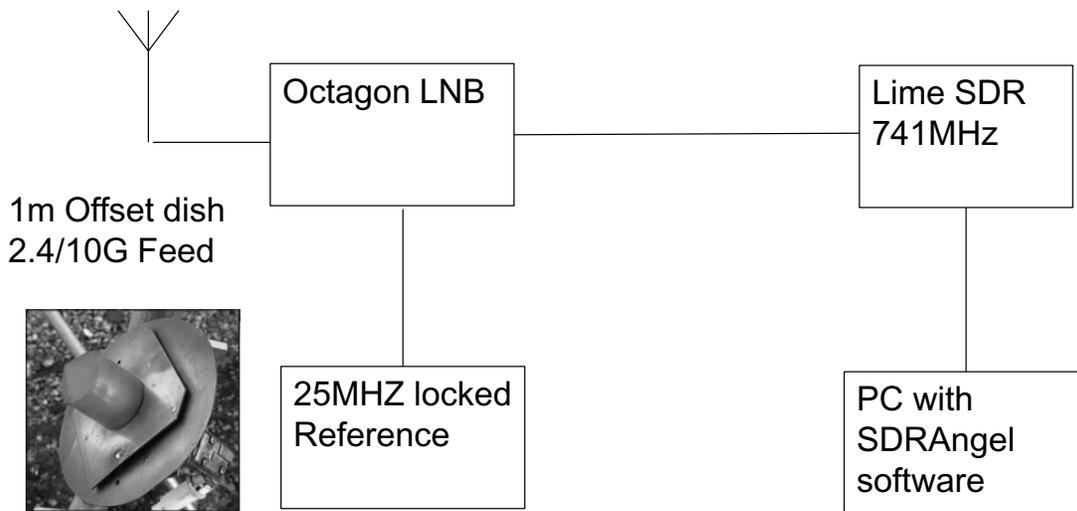
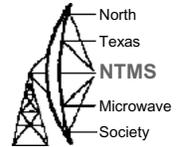
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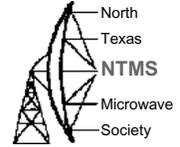
22:44 fSoeo_evariste Waht is your GOP size ?
22:45 fSoeo_evariste Seems very short..12 frame or something
22:45 PE2JKO_Joop 100
22:46 fSoeo_evariste Strange, not sure about 100...seems very short (inpecting timing)
22:46 G7JT-John Big delay
22:46 G7JT-John Yes have sound
22:52 fSoeo_evariste As I understand, RTMP is the "over the edge" technic ...
22:54 fSoeo_evariste You are now workaround my lime encoder bugs ..specially buffering... You are great !
22:57 fSoeo_evariste Joop do you use FPGA mode (lot of latency now) ?
22:58 fSoeo_evariste OK, you have to do the clown now John to see...
23:00 PE2JKO_Joop no FPGA mode, due to the fact that I also use the same for SSB
23:00 G7JT-John I will try your software tomorrow Evariste and see how that works as well
23:00 PE2JKO_Joop I use Gnuradio + GR-CESB which can't work with FPGA mode
23:01 fSoeo_evariste OK Joop, we need to implement a pass-through mode with the firmware to avoid that
  
```

Type '/nick your_name' and press enter to register.

<https://eshail.batc.org.uk/wb/>

TV Downlink Receiver





Calculating how much uplink power is needed

BATC **How much power?**

Dish Diameter	Power Required for 250Ks
2.4m	7.5W
2.0m	11W
1.8m	13W
1.5m	19W
1.2m	30W
1m	43W
90cm	53W
80cm	67W
60cm	120W

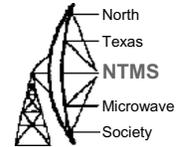
Power required to achieve same MER as the beacon at a receiving station – eg 8dB MER on 1.2mt dish

SR	Factor
66 KS	0.26
125 KS	0.5
250 KS	1.0
333 KS	1.33
500 KS	2.0
1000 KS	4.0
1500 KS	6.0

125Ks = 15 watts with 1.2mt dish for 8dB MER

Relative Power	Received MER
100%	8 dB
80%	7 dB
63%	6 dB
50%	5 dB
40%	4 dB
31%	3 dB
25%	2 dB
20%	1 dB
16%	0 dB
13%	-1 dB
10%	-2 dB

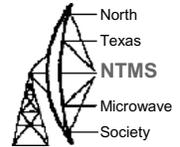
7.5 watts, 1.2mt dish = 125Ks at 5dB MER



DATV Microwaves

Simple as can use Portsdown to drive Microwave Narrowband Transverter

24GHz DATV

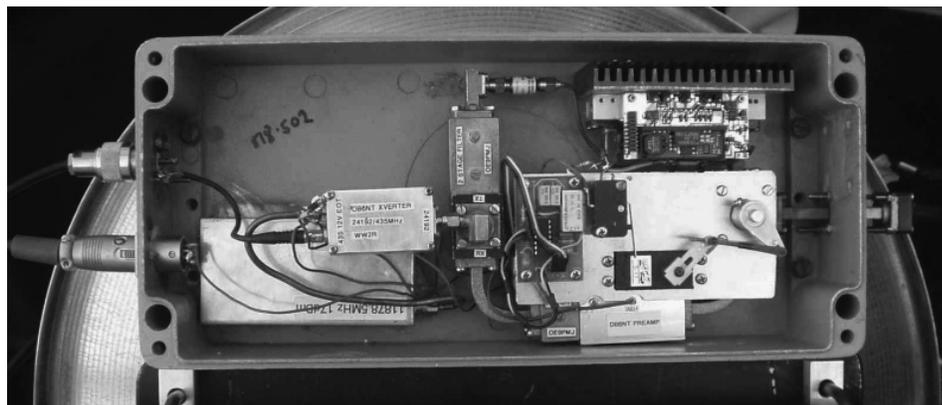
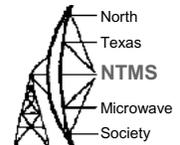


World record was 122km in Japan
 First UK attempt was Brown Clee (IO82QL) to
 Winter Hill (IO83SO) 126.6km



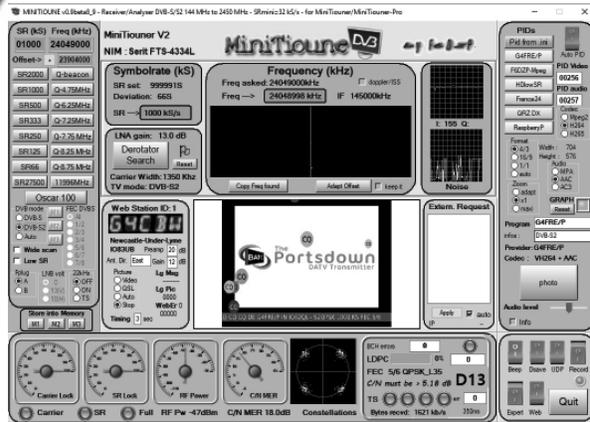
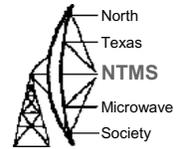
G4FRE/P
 IO82QL

My 24GHz DATV Transverter



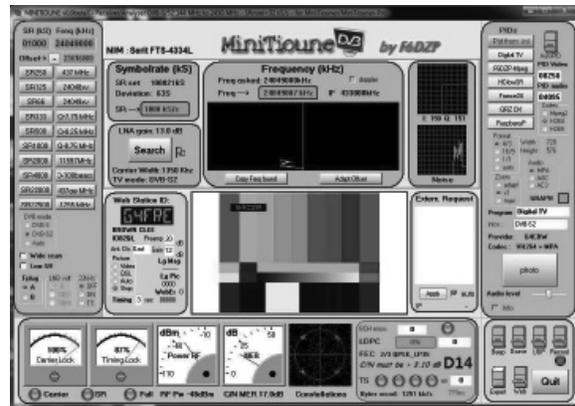
0.5W DATV 1.8dB noise figure

24GHz DATV 126.6km

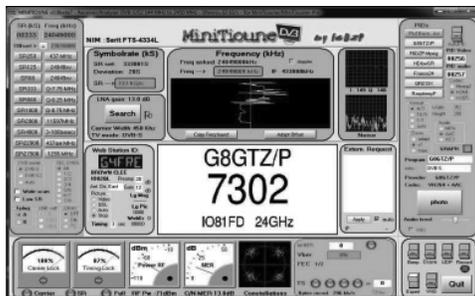
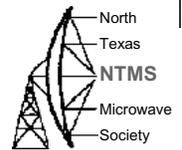


<<<<G4FRE/P received at G4CBW/P

G4CBW/P received >>>> at G4FRE/P

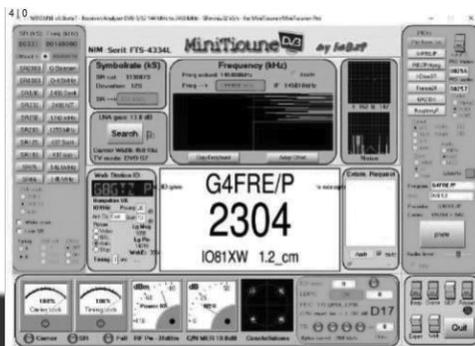


Three weeks later during IARU TV Contest



G4FRE/P IO81XW

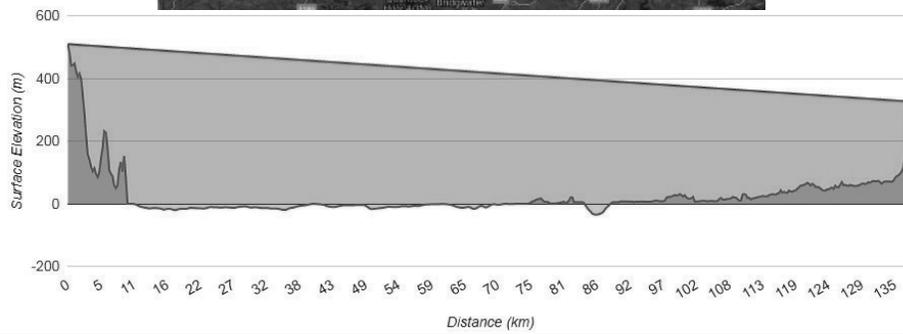
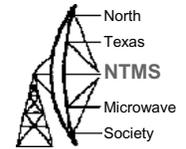
136km



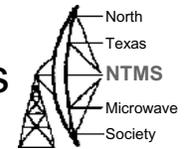
G8GTZ/P IO81FD



Cleeve Common to Dunkery Path



Other Microwave Band DATV Records



10GHz

407 kms	M0DTS/P	IO94MJ	G4UVZ	IO80KX	2Ms/s
---------	---------	--------	-------	--------	-------

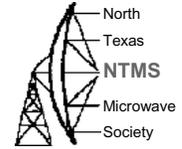
47GHz

34.9 km	G4FRE	IO82UC	G0FRE/P	IO92BA	1 Ms/s
---------	-------	--------	---------	--------	--------

76GHz

35.6 kms	G8GTZ/P	IO91GI	G4LDR/P	IO91JB	333Ks/s
----------	---------	--------	---------	--------	---------

What about lower Bands?



71MHz

UK full amateurs can get NOV on 70.5-71MHz for “experimental modes”

Using Portsdown + G4DDK Nacton 437/71MHz XV

160 Kms	G8GTZ/P	IO80WX	G4FRE/P	IO82QJ	125Ks
---------	---------	--------	---------	--------	-------

144MHz

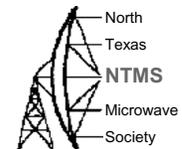
UK full amateurs can get NOV on 145-146MHz for “experimental modes”

407 kms	M0DTS/P	IO94MJ	G4UVZ	IO80KX	333Ks
---------	---------	--------	-------	--------	-------

Note narrower bandwidths

140km also worked on 51MHz!

References



<https://batc.org.uk/>

<https://www.crowdsupply.com/lime-micro/limesdr>

<https://www.crowdsupply.com/lime-micro/limesdr-mini>

<https://wiki.analog.com/university/tools/pluto>

<https://github.com/f4exb/sdrangel>

<https://www.sdr-radio.com/Software/>

<https://eshail.batc.org.uk/nb/>

<https://eshail.batc.org.uk/wb/>

Using the TI LMX2594 Synthesizer for Microwave Local Oscillators

If you have been involved in microwave experimentation for more than a decade then you probably remember the days of trying to generate a clean microwave local oscillator using a crystal oscillator in the VHF range and then multiplying it using various bipolar RF transistors. It was the only way to generate reasonably stable LO's but the downside was the difficult tuning of the stages and all the stages that also like to oscillate by themselves plus it was also impossible to lock them to a more stable reference when dealing with very low bandwidth signals (CW and SSB). There was available many versions of microwave surplus cavity oscillators that were locked to a VHF crystal operating around 100MHz and a high power cavity oscillator would drive a multiplier to output a multiple of the crystal but these too were tweaky and ran hot on usually -19 VDC at a lot of current and also were very microphonic. About a decade ago Hittite Microwave (now part of Analog Devices) came out with highly integrated synthesizers that contained the VCO, buffers and support circuitry to phase lock a low frequency reference signal using either integer N or fractional N dividers. They also integrated onto the IC the ability to divide the VCO output frequency to generate a very wide range of LO frequencies. Over the last few years there has been quite a few of these ICs developed and one of the latest synthesizer ICs is the Texas Instruments (TI) LMX2954.

The LMX2954

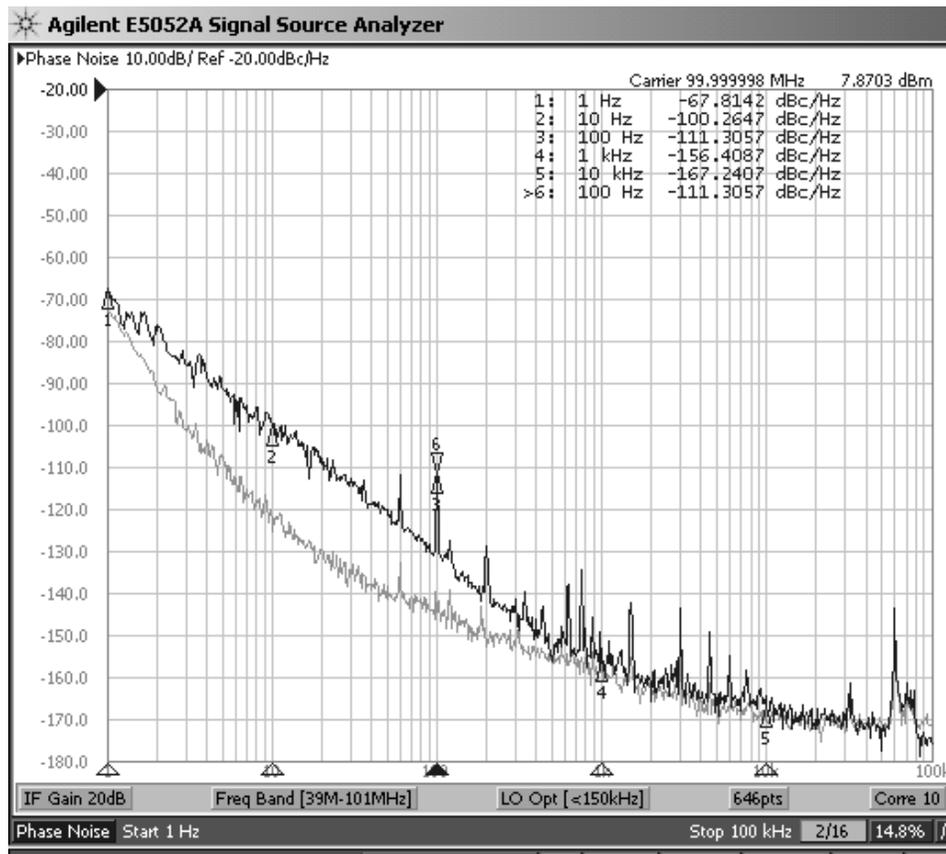
This IC is a complete synthesizer on a chip, it includes almost everything needed to generate an LO from 10MHz to 15GHz with very good phase noise if a suitable reference oscillator is selected. The IC is very complex and its way beyond the scope of this paper to even touch on all the features but TI has an excellent set of technical documents available online and also offers an evaluation unit on a PC board with a USB interface for designers to easily fabricate a microwave LO. The evaluation board is easy to get operational but it's not designed to incorporate inside your up or down converter unless you plan on connecting to a PC every time you power the unit up. The USB interface also contains a low phase 100MHz reference oscillator but this unit does not have the necessary stability to use for narrowband operation such a CW or SSB. While the evaluation board is fun to experiment with, in order to really use the device a new board needed to be designed with the ability to store the synthesizer register values so the unit would power up ready to be used, it can then be incorporated into portable equipment for field operation. For our application we interfaced the LMX2594 to a PIC microprocessor thru the SPI interface and then loaded a program via a JTAG interface where the PIC microprocessor would store the desired LO frequency and set up all the registers. Since the synthesizer requires a reasonably clean 3.3 VDC we also used the TI very low noise linear regulator which was driven by a wide input range switching power supply to improve the power supply efficiency and reduce the heat from dropping a large voltage across the linear regulator. The LMX2594 has internal low dropout regulators included with the IC but it's still good practice to supply the IC and especially the reference oscillator with low noise DC linear regulators.

This cannot be overstressed, the DC must be clean and free of any noise or it will degrade the overall synthesizer phase noise. The DC connection to the reference oscillator and LMX2594 should also be with short leads that are routed away from any switching power supplies or else you will start to see

spurs at the switching frequency on the LO output. Its also not uncommon to see 60Hz and 120Hz low level spurs on unshielded LO's as you will note in some of the following phase noise plots of the LMX2594 reference oscillators as the E5052A phase noise analyzer is capable of measuring extremely low levels of phase noise and spurs.

Reference Oscillators

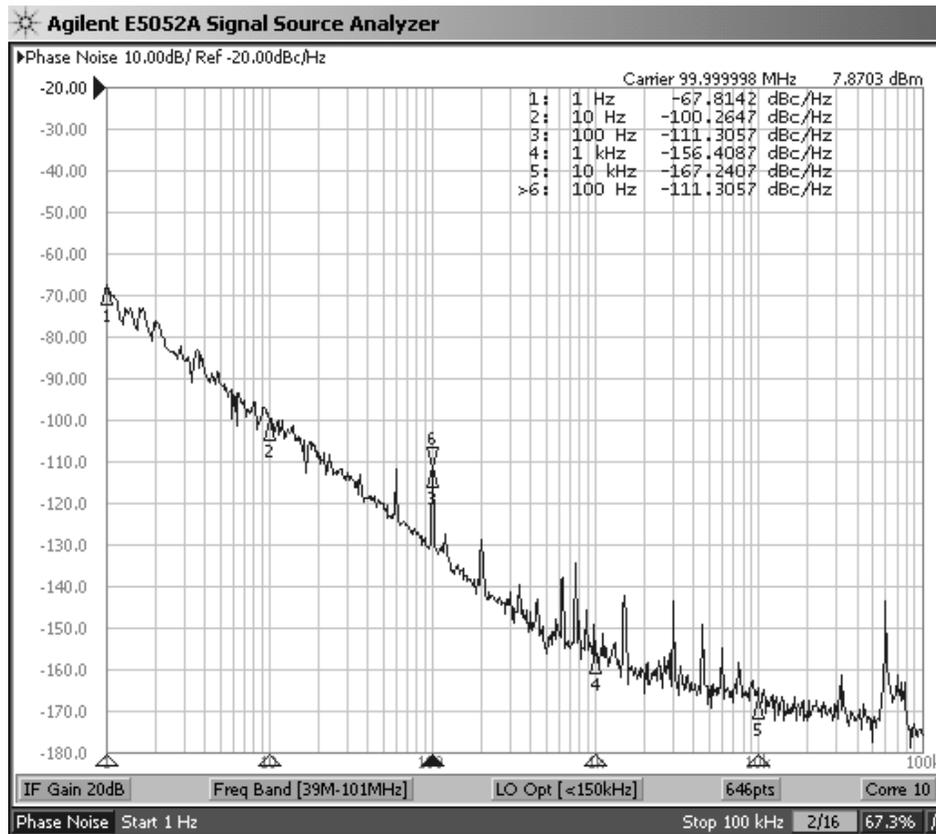
Synthesizers are only as good as their reference oscillators. Synthesizers take on the phase noise characteristics of the reference oscillators phase noise in the loop bandwidth of the synthesizer but degraded at $20 \log F_{out}/F_{ref}$. As an example if we start with a 10MHz reference oscillator instead of a 100MHz reference oscillator we take a $20 \log 100\text{MHz}/10\text{MHz}$ or 20dB worse to start off with than if we started with a 100MHz reference oscillator. This is why these days most high performance microwave synthesizers start with a low phase noise 100MHz reference oscillator or at least higher than 10MHz to reduce the phase noise in the loop bandwidth at the output frequency. Typically the 100MHz phase noise reference oscillator won't have as good as a 10MHz reference oscillator but its still better than starting with 10MHz unless you are synthesizing a low frequency LO.



008 Plot 1 10MHz OCXO Reference Oscillator and 100MHz OCXO Reference Oscillator Compared Phase Noise

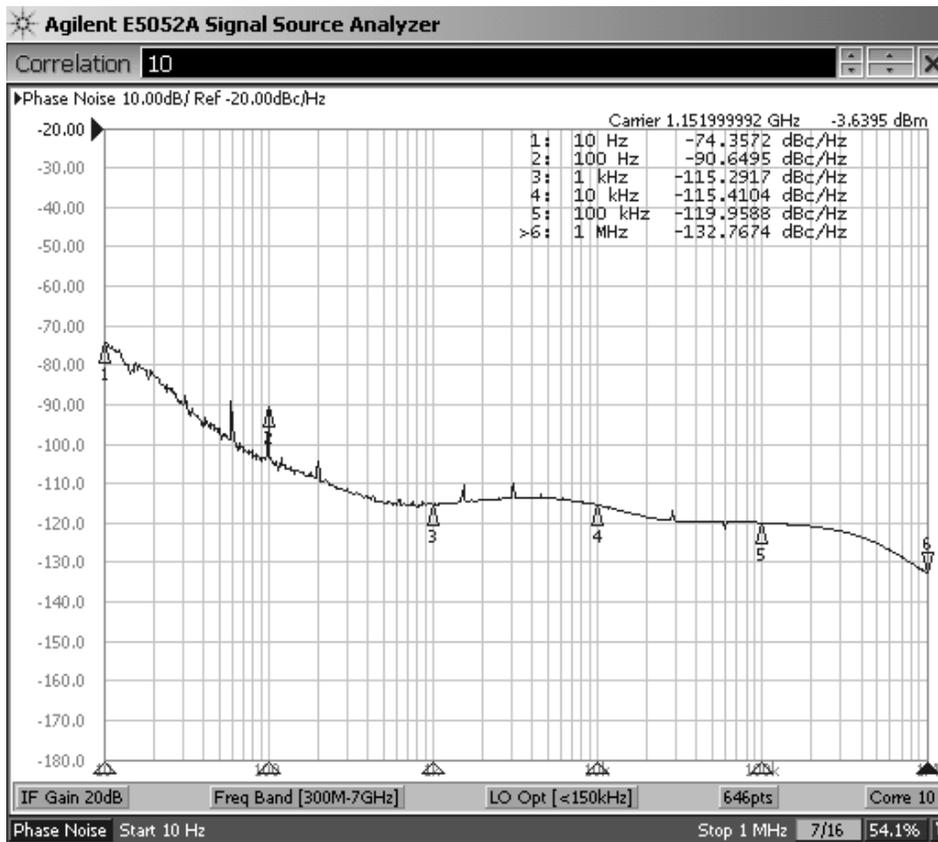
Plot 1 shows two reference oscillators, one at 10MHz and one at 100MHz. The lower phase noise plot is the 10MHz OCXO, the initial phase noise is about the same at 10Hz but the 100MHz is not as good until you reach around 1kHz. Remember that the 10MHz reference oscillator by the time its compared

at 100MHz will degrade 20dB. These two oscillators are high performance OCXO units made by Wenzel Associates, the 100MHz OCXO during these tests was locked with a loop bandwidth of <5Hz to a 10MHz GPS reference as otherwise it could be off frequency as much as 10kHz, the 10MHz OCXO used in the test was not phase lock but was a free running OCXO. These are exceptionally good reference oscillators and are pricey but I do see them available along with other good oscillators available as surplus from time to time on Ebay.



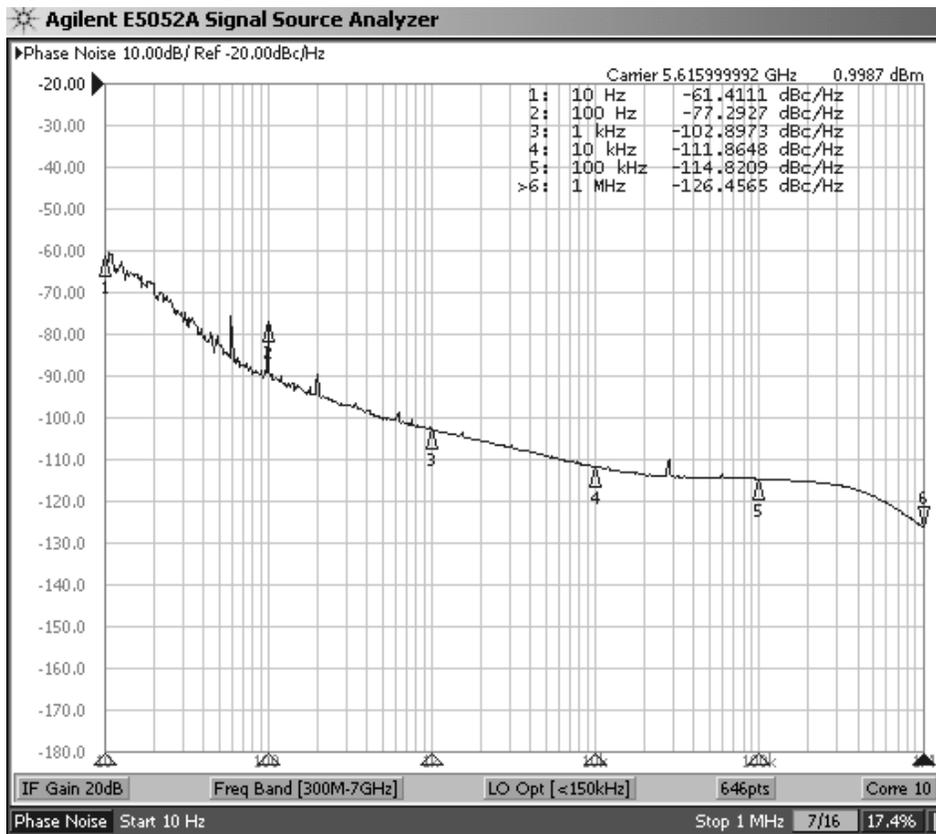
009 Plot 2 100MHz Wenzel Reference Oscillator locked to GPS 10MHz Reference Signal

Plot 2 Shows a good quality low phase noise OCXO that is locked to a GPS 10MHz reference signal with the GPS locking PLL with a loop bandwidth of about 5Hz. Most of the following phase noise plots use this reference oscillator to drive the LMX2594 Eval PCB using a single ended 50 ohm drive signal at about +6dBm.



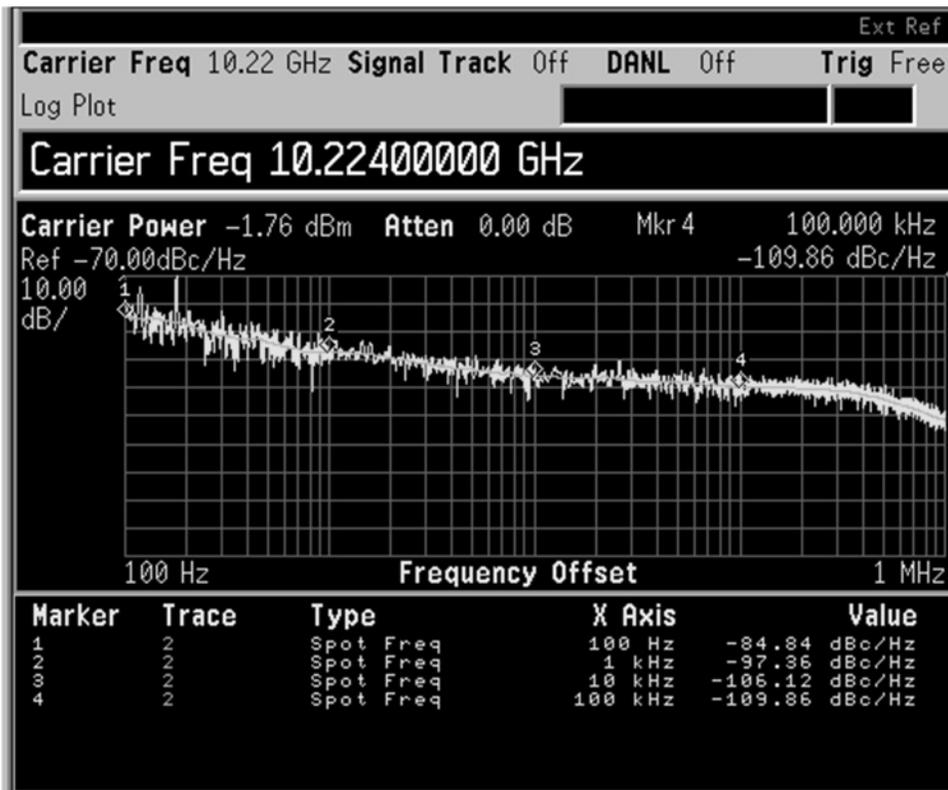
005 Plot 3 1152MHz LO for 1296MHz converter phase noise

Plot 3 shows the LMX2594 Eval board set up to generate an LO for a 1296MHz converter using a 144MHz IF difference. The LMX2594 VCO is operating at 9216MHz and the output is being divided down by 8 to the final 1152MHz. By dividing by 8 the actual oscillators phase noise is being improved by $20 \log 8$ or 18dB. These phase noise values are pretty good for this type LO, only a phase locked DRO would be better.



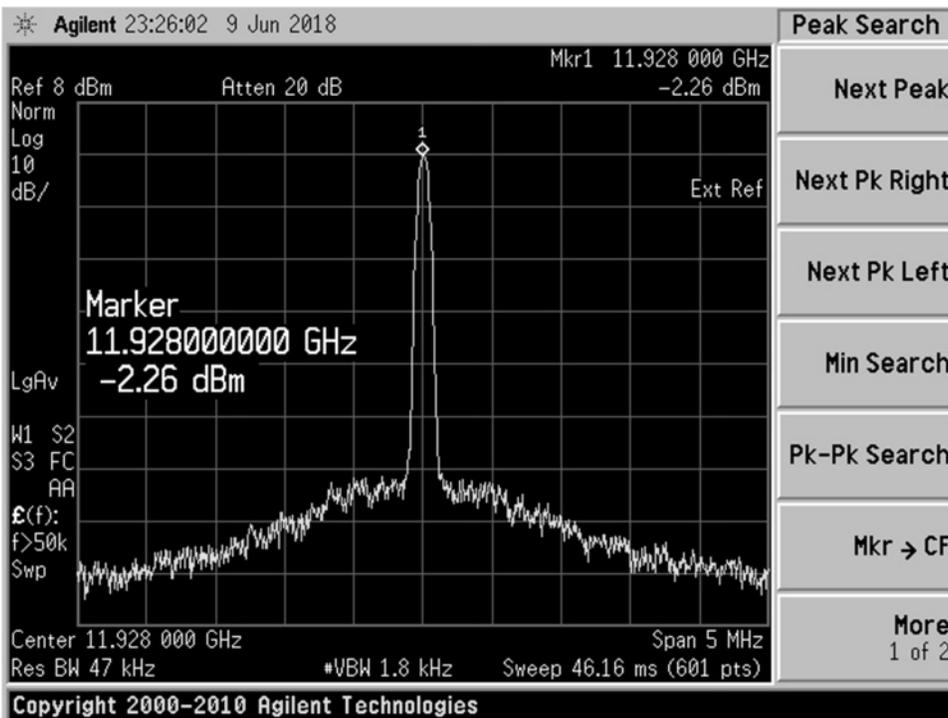
004 Plot 4 5616MHz LO for use with a 5760MHz transverter phase noise plot

Plot 4 shows the eval board reprogrammed to operate at 5616MHz using the 100MHz reference oscillator. The LMX2594 reference input frequency is being doubled by the IC to 200MHz and the phase detector is being operated at 200MHz. The LMX2594 VCO is operating at 11232MHz and the output has been programmed to divide the VCO frequency by 2. The phase noise is still very good, the output phase noise is being improved by 6dB by the divide by 2 on the output.

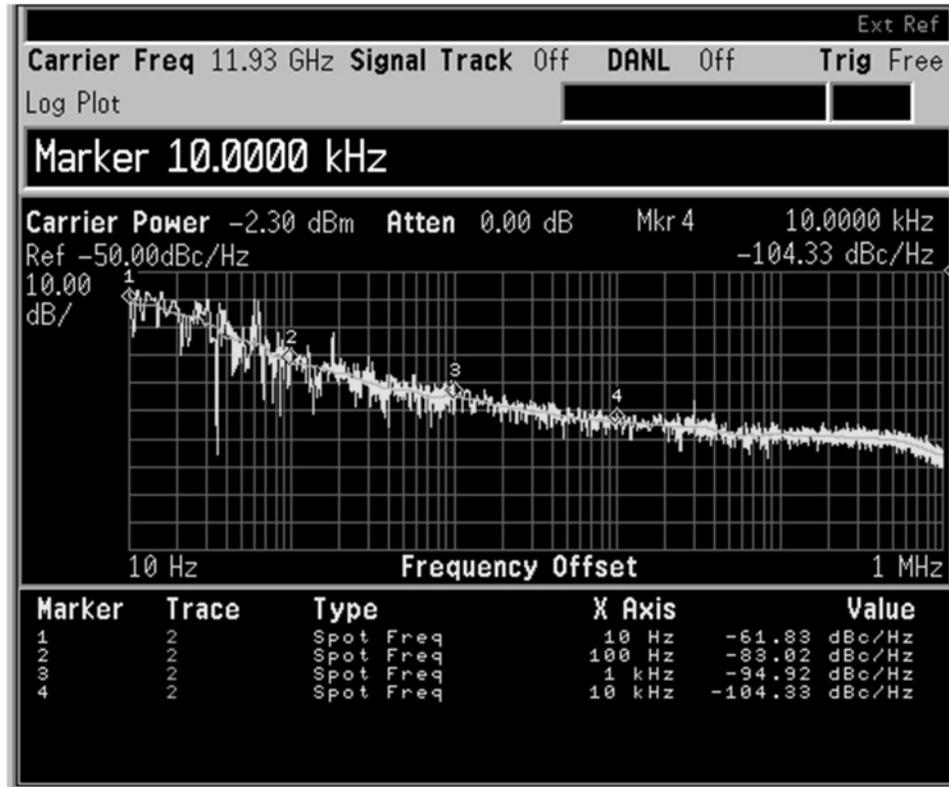


SCR513 10224MHz LO for 10368MHz transverter using a 144MHz IF

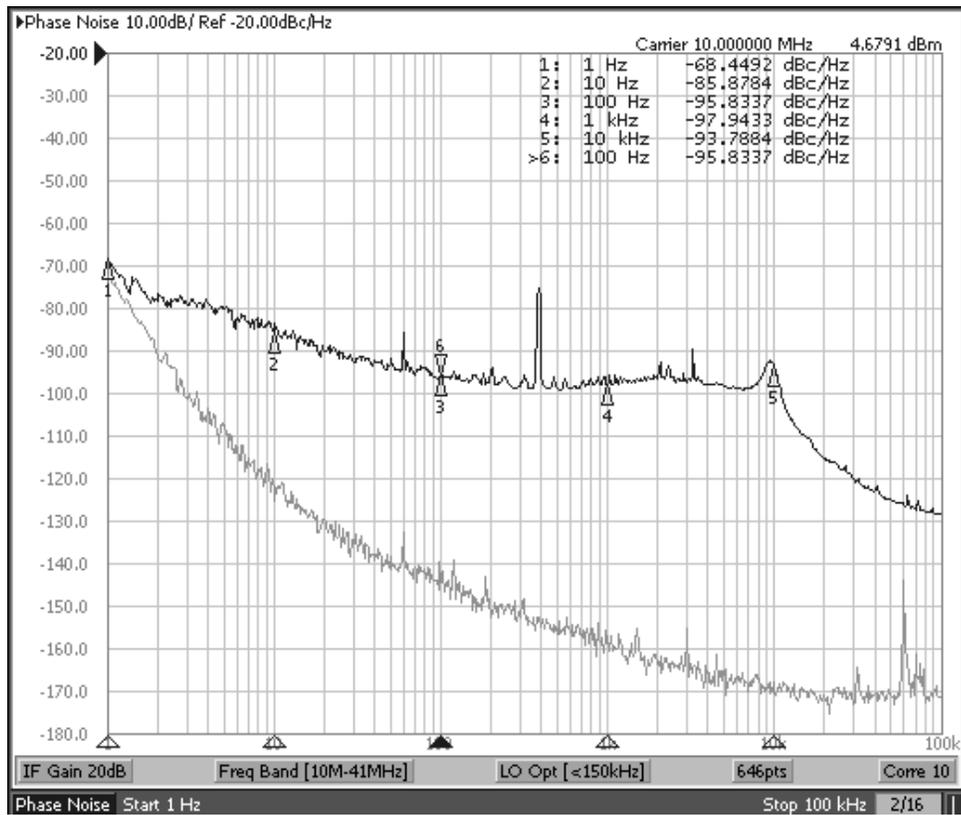
This phase noise plot is from a spectrum analyzer since the E5052A Signal Analyzer tops out at 7GHz. The phase noise plots using the spectrum analyzer will not be as accurate as the E5052A but are close enough to see the general trend in the LMX2594 as the LO frequency goes higher.



SCR514 Reprogrammed to generate a 11928MHz LO which can be used with a subharmonic mixer to make a 24GHz transverter. The LO operates at $\frac{1}{2}$ frequency and is doubled in the subharmonic mixer, this alleviates the need to double the LO frequency before the mixer. Keep in mind that doubling the LO frequency in the subharmonic mixer will degrade the phase noise by 6dB from that shown on the plot.

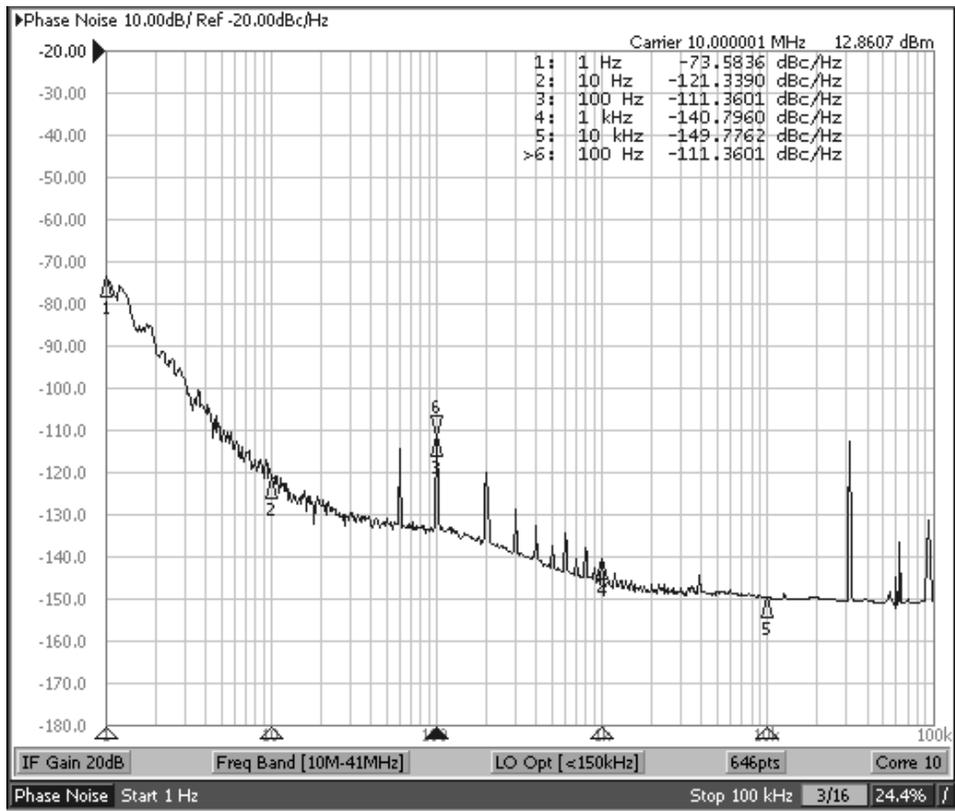


Scr 515 Phase noise plot for the 11928MHz LO



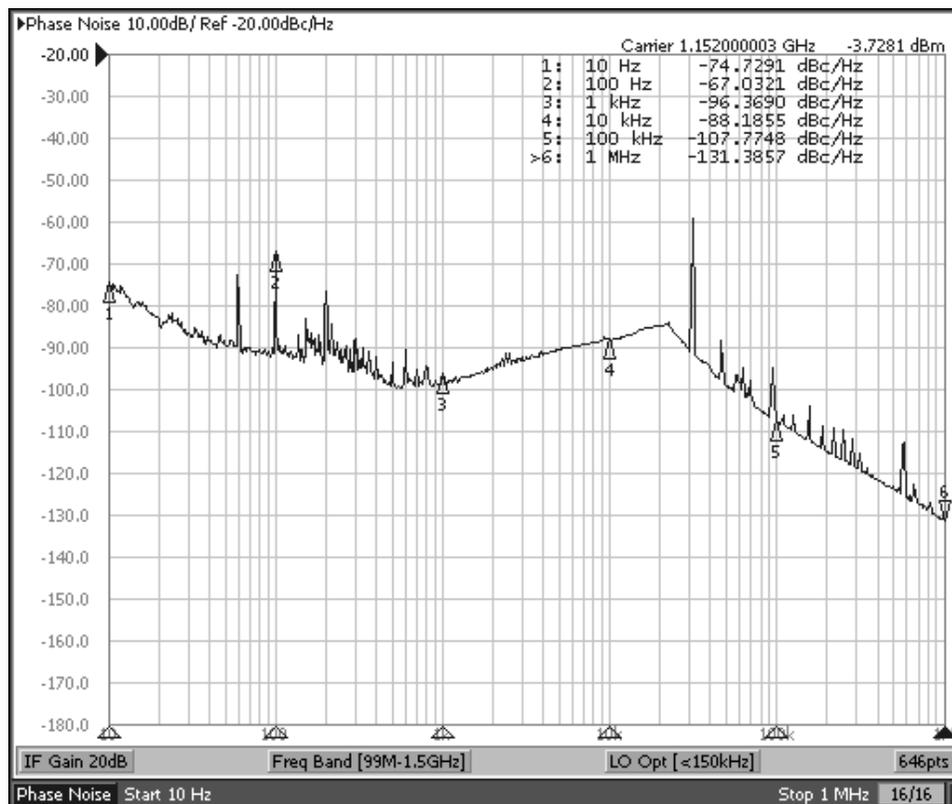
0010 Comparison of a Wenzel 10MHz free running OCXO vs 10MHz GPS phase noise

While making the measurements I decided to also show several extra plots of different oscillators and how they compared to each other. The plot above shows a Wenzel 10MHz OCXO free running reference oscillator (bottom trace) and a 10MHz output from a typical signal generator. This plot clearly shows why you never use a signal generator for a synthesizer reference unless just testing a circuit.

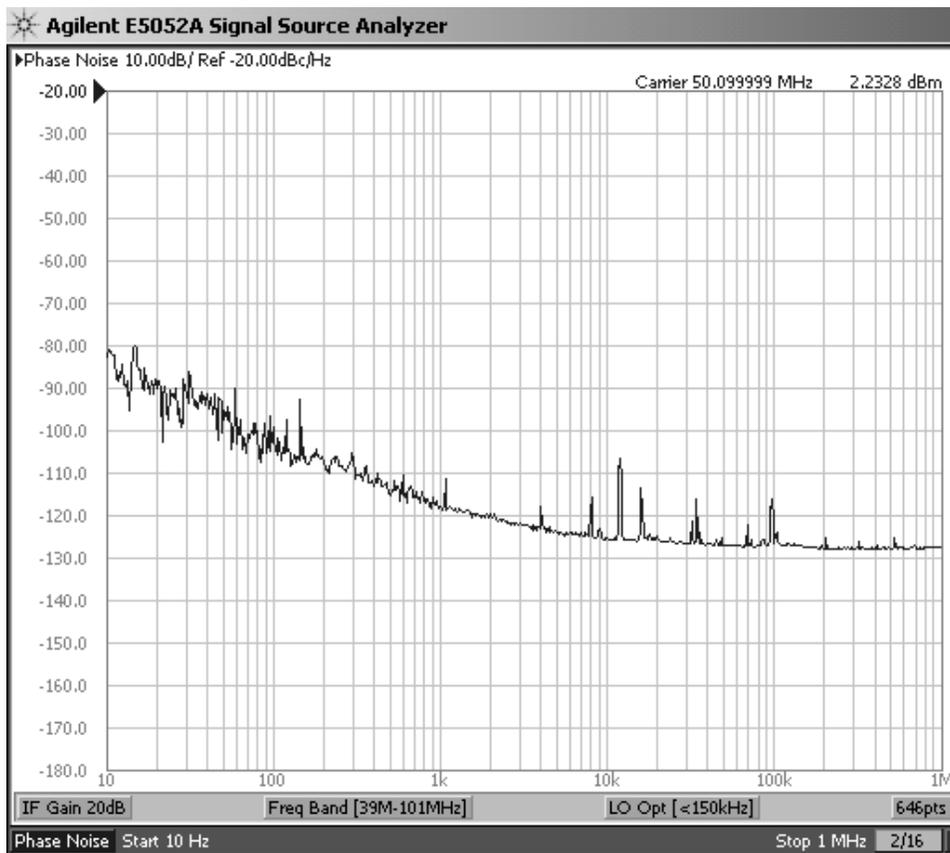


0013 Phase Noise plot of a typical 10MHz GPS reference signal used to lock a stations other oscillators.

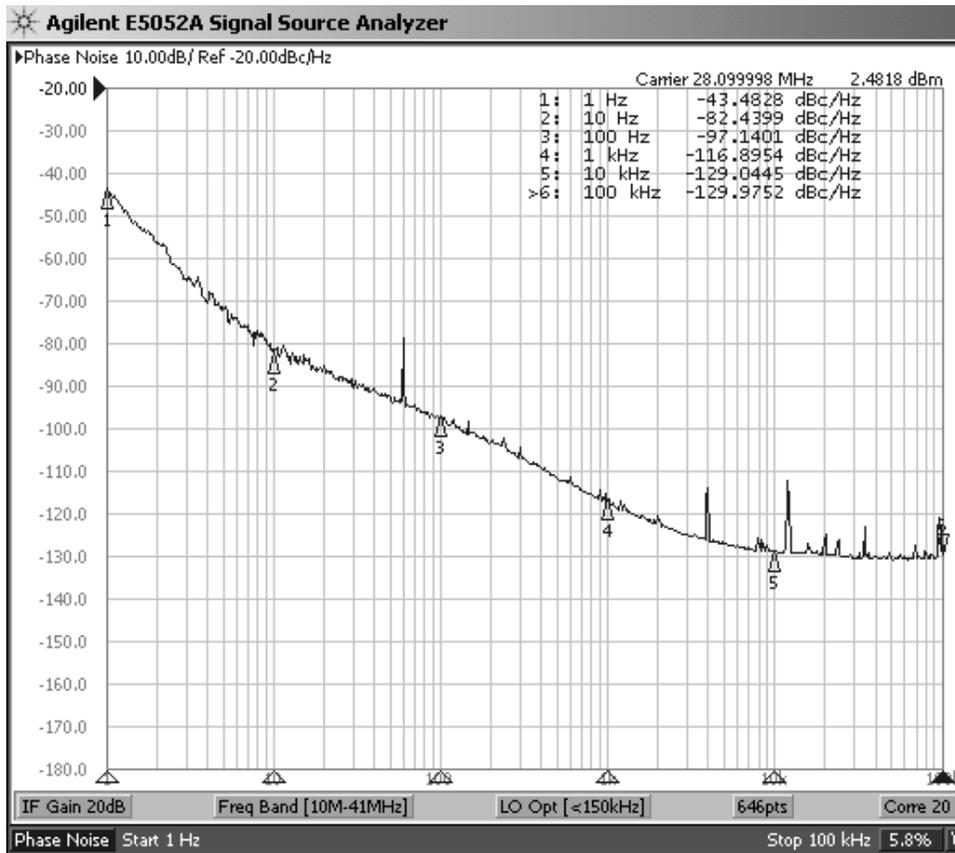
This plot shows the phase noise out of a typical GPS 10MHz disciplined reference signal that is distributed to other LO's to lock the frequency to a precision frequency reference. This signal is not intended to be used to as a synthesizer reference oscillator as I will show in the next plot. Most synthesizers that take an external 10MHz reference use a PLL with just a few Hz bandwidth to lock the frequency up but not to change the synthesizer phase noise, that is done by the local reference oscillator.



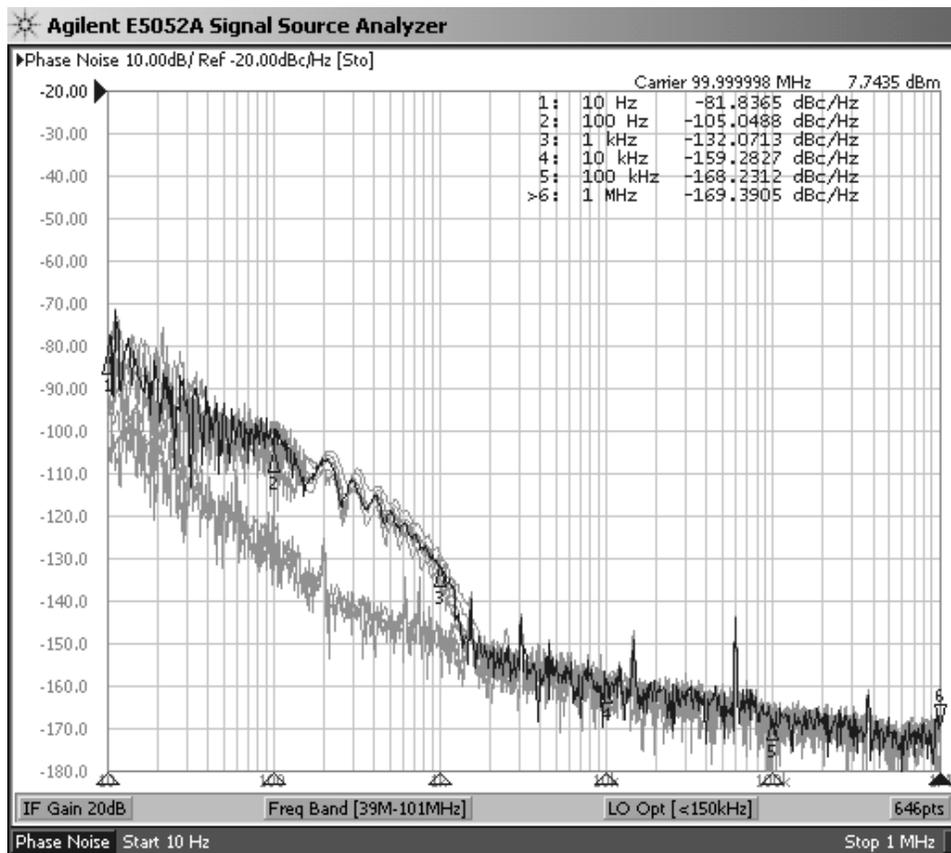
0015 1152MHz LO using the GPS 10MHz only as the reference oscillator. While this would probably work its certainly not a clean LO by today's standards.



ICOM-7300 Transceiver 10Hz to 1MHz phase noise 50.1MHz transmit frequency set to 20% power output with a 40dB power attenuator in line. These transceivers offer great performance and very clean signals as compared to the early radio synthesizers. Its good to keep in mind that this clean transmitter and receiver will be degraded if you mix these signals in a transverter with poor phase noise performance. Your mixed signals will take on the characteristics of the poor synthesizers phase noise.



ICOM-7300 Phase Noise 28.1MHz 20% power output CW. This plot extends from 1Hz to 100kHz.



Plot showing the 100MHz Reference Oscillator with light tapping on the case. Most reference oscillators are microphonic, usually much more so than the monolithic VCOs now used in the IC synthesizers. This plot shows the degradation between 10Hz and 2kHz of over 30dB due to mechanical vibration. When used as a reference oscillator in a microwave LO these hits can cause the synthesizer to unlock or at least cause the received signal in your transverter to sound awful. If you are really trying for the highest quality LO's you will probably need to shock mount at least the reference oscillator. Some oscillators are much more sensitive to vibration in one axis than the other due to the way the crystal is mounted. Reference oscillator vendors are becoming more aware of this and are working on oscillators that are less microphonic. Bliley has introduced some tiny TCXO's operating up to 50MHz that while not having as good phase noise performance of the best OCXO they are coming close to this performance with very low microphonics. Using one of these as a reference oscillator will give you around -95dBc/Hz performance at 10kHz offset at 5GHz.

The LMX2594 and other monolithic synthesizers are amazing devices as compared to the way we had to generate LO's even a decade ago. The designers have packed an fantastic amount of flexibility into these devices to lower the noise and coupled with a good reference oscillator, a small microprocessor and a clean DC supply today we can generate about any frequency desired with few spurious signals.

DESIGN FOR AN AMATEUR RADIO SATELLITE COMMUNICATIONS ANTENNA

Donald Shea
Applied Antenna Technology
Allen, Texas

Abstract

This paper describes the design for a parabolic antenna system which can be used for communications with GEO satellites at the assigned amateur radio frequencies in S-Band and X-Band. A coaxial waveguide feed is used in this design which provides excellent performance in both bands. The concept for this type of feed is shown in Figure 1.

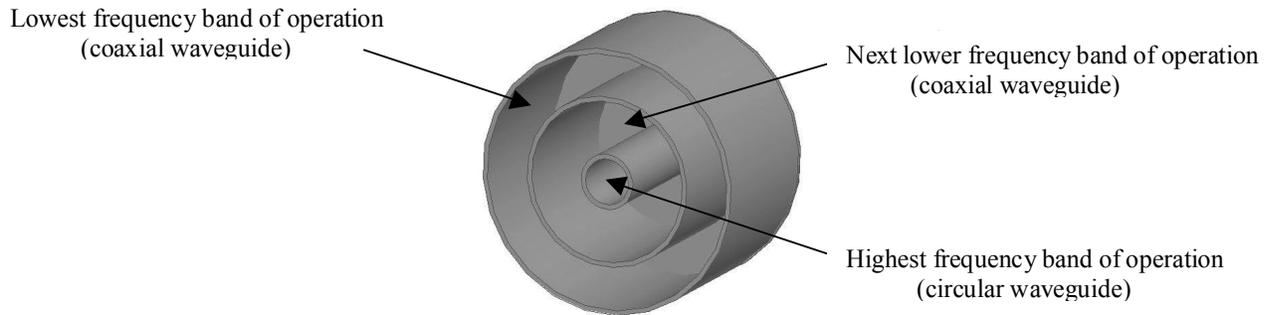


Figure 1. Coaxial feed configuration for multi-band operation

Background

Coaxial waveguide was analyzed and described in the M.I.T. Radiation Laboratory Waveguide Handbook, Vol. 10, published in 1950; however, subsequent to this publication, use of coaxial waveguide as an antenna feed has not been common. Its earliest use was in the design of a dual band (VHF/UHF) high power radar system deployed at White Sands Missile Range in 1963 shown in Figure 1.



Figure 1. VHF/UHF radar system

It was recognized that this new approach was superior to the multi-band feed systems then deployed and showed that a parabolic reflector with a multi-band coaxial feed could perform the work of several reflectors using single band feeds without sacrificing performance.

A mobile tri-band SATCOM system, using a coaxial waveguide feed (Figure 2), was developed for the U.S. Army on an SBIR contract in 1999.



Figure 2. Tri-band SATCOM antenna

Antennas of this type were subsequently produced and deployed for both commercial and military SATCOM applications. Examples are shown below:



Large Tri-band SATCOM Ground Station



Transportable Tri-band SATCOM Antenna



Tri-band Programmed Track Antenna

Principle of Operation

To function as a prime focus feed, circular waveguide must operate in the $TE_{1,1}$ mode. The field distribution for this mode is shown in Figure 3.

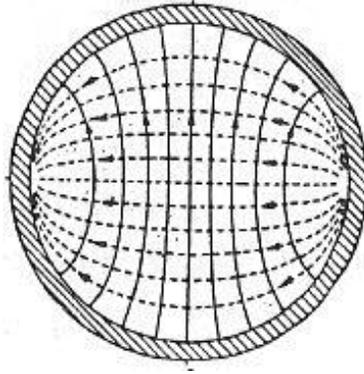


Figure 3. Electro-magnetic fields in circular waveguide for $TE_{1,1}$ Mode

$TE_{1,1}$ propagation in this waveguide is possible when the cavity diameter is $> 0.6\lambda$.

Likewise, coaxial waveguide must be operated in the $TE_{1,1}$ mode. The field distribution for this mode is shown in Figure 4.

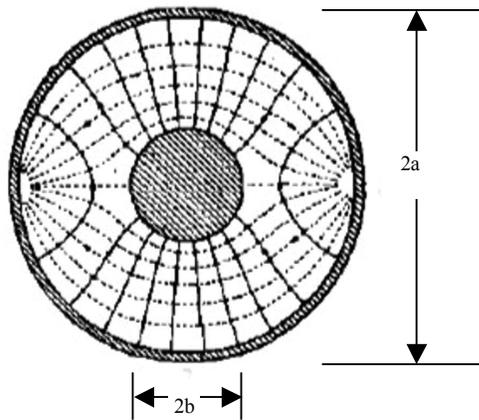


Figure 4. Electro-magnetic fields associated with $TE_{1,1}$ mode in coaxial waveguide

$TE_{1,1}$ propagation is possible when the mean circumference is greater than 1 wavelength. The cutoff wavelength is therefore given by the formula:

$$\lambda_{co} = \pi(a+b)$$

An open-ended waveguide of this type radiating into space produces an on-axis pattern as shown in Figure 5. Quadrature phased orthogonal $TE_{1,1}$ modes can be simultaneously excited in the waveguide to produce circular polarization.

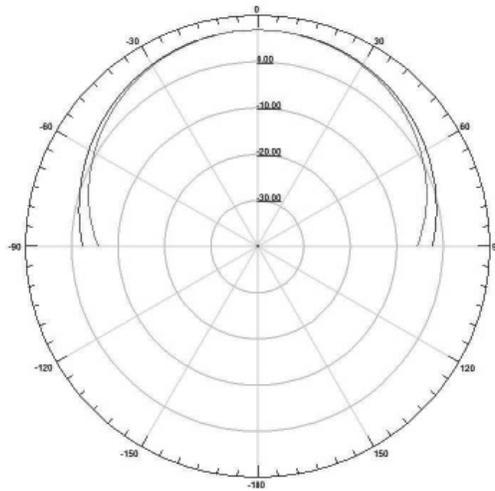


Figure 5. $TE_{1,1}$ Pattern produced by open-ended coaxial waveguide

As seen in this figure, the pattern produced by open ended coaxial waveguide is near optimum for use in a parabolic reflector.

Antenna Design

The design shown in Figure 6 provides the desired performance while being a cost effective, conventional approach. It consists of an OEM parabolic reflector with a prime focus, dual band coaxial feed.

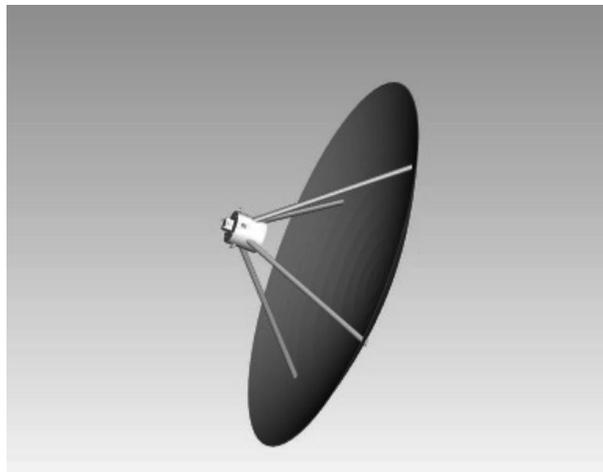


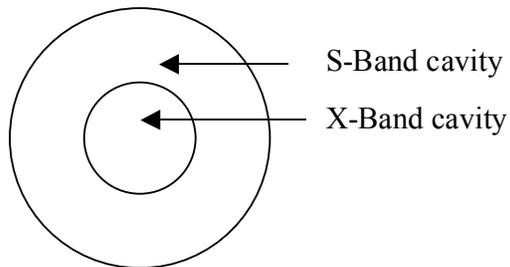
Figure 6. Antenna system for amateur band satellite communications

The feed is designed to operate over the following frequencies:

S-Band: 2404.775 MHz \pm 4.725 MHz
X-Band: 10,494.275 MHz \pm 4.725 MHz

Polarization shall be RHCP (from the reflector) for S-Band and dual linear polarization for X-Band.

It is implemented with two concentric coaxial waveguide cavities arranged as shown below.



The inner cavity is a conventional open ended circular waveguide covering X-Band; the outer is coaxial waveguide covering S-Band.

The outer cavity is fed with four orthogonally placed probe transitions as shown in Figure 7. The feed excitation network to produce circular polarization is shown in Figure 8.

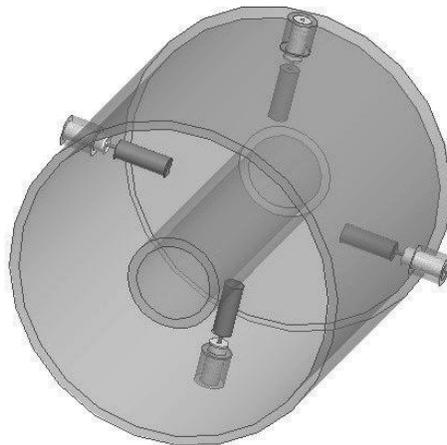


Figure 7. S-Band feed excitation

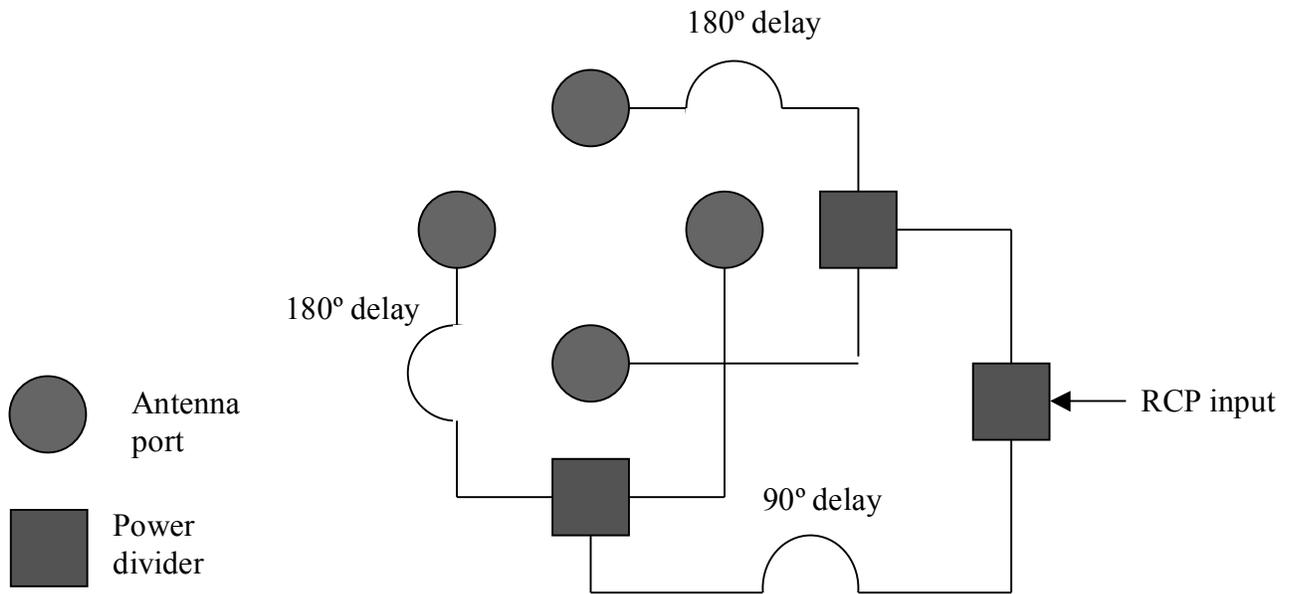


Figure 8. Network for producing circular polarization

The inner (X-Band) waveguide cavity is fed with dual probes as shown in Figure 9.



Figure 9. X-Band feed excitation

HFSS (High Frequency Structure Simulator) software was used to design and evaluate the feed shown in Figure 10.

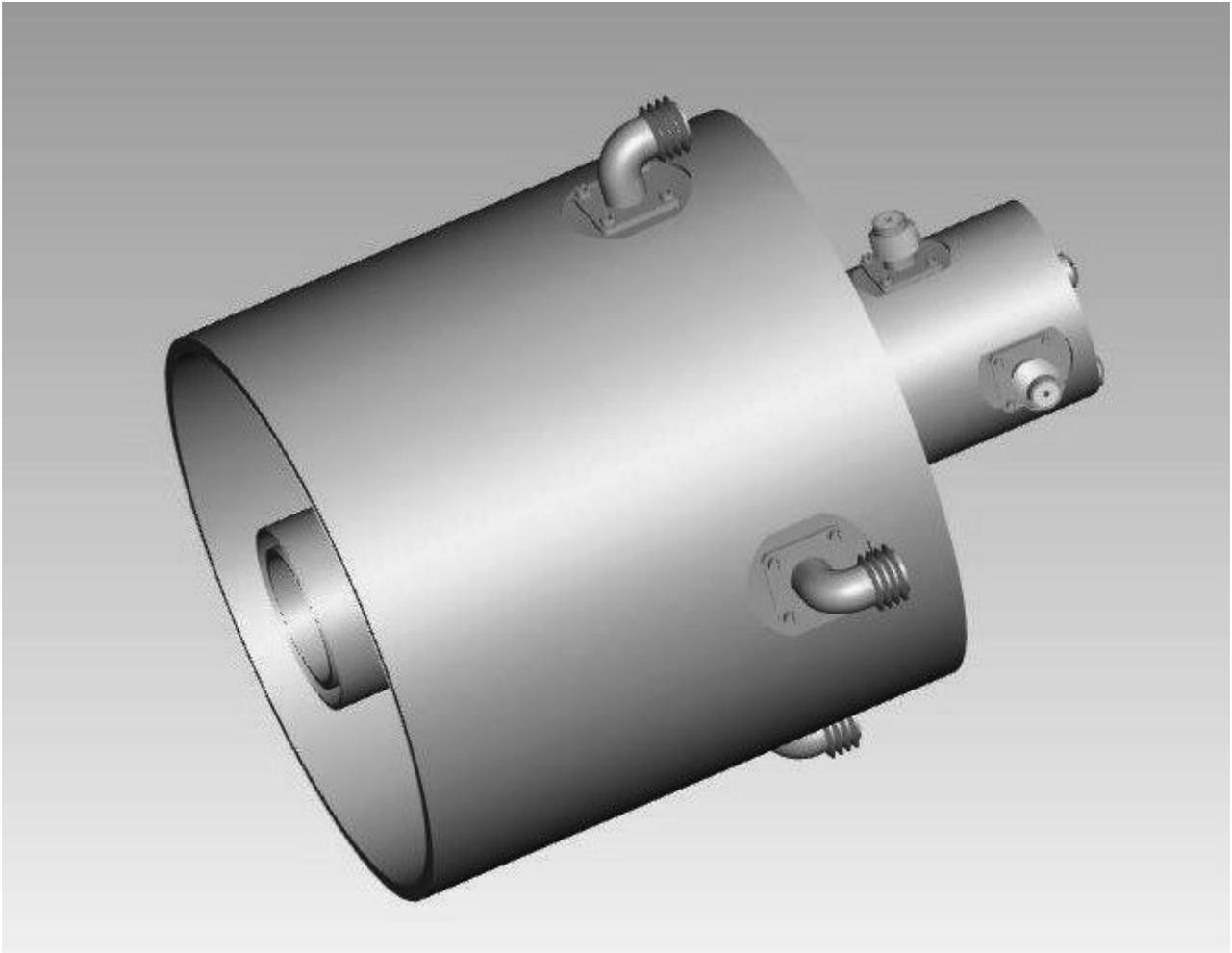
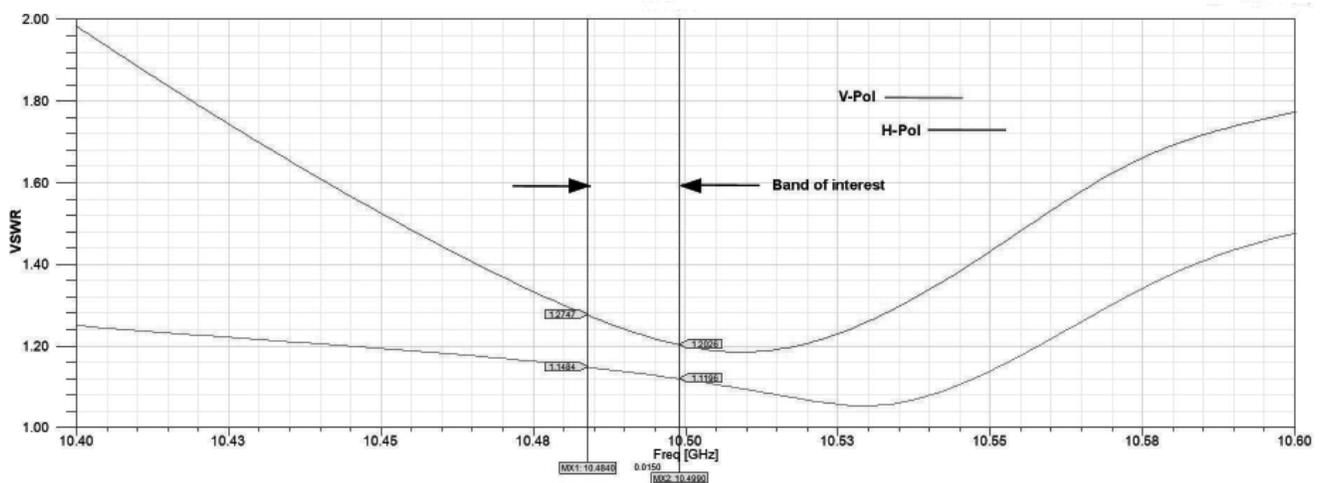
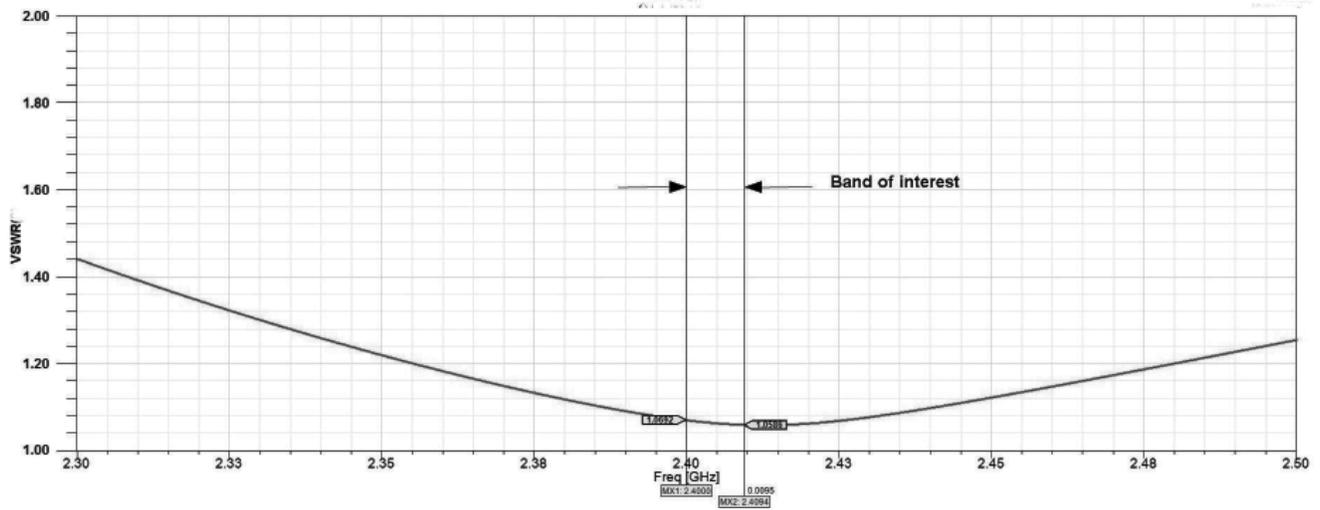


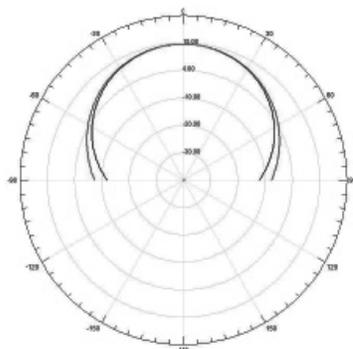
Figure 10. S/X-Band feed

Computed Performance

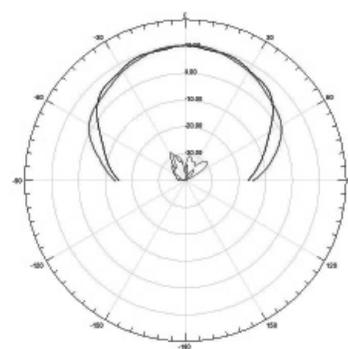
VSWR



Primary Patterns

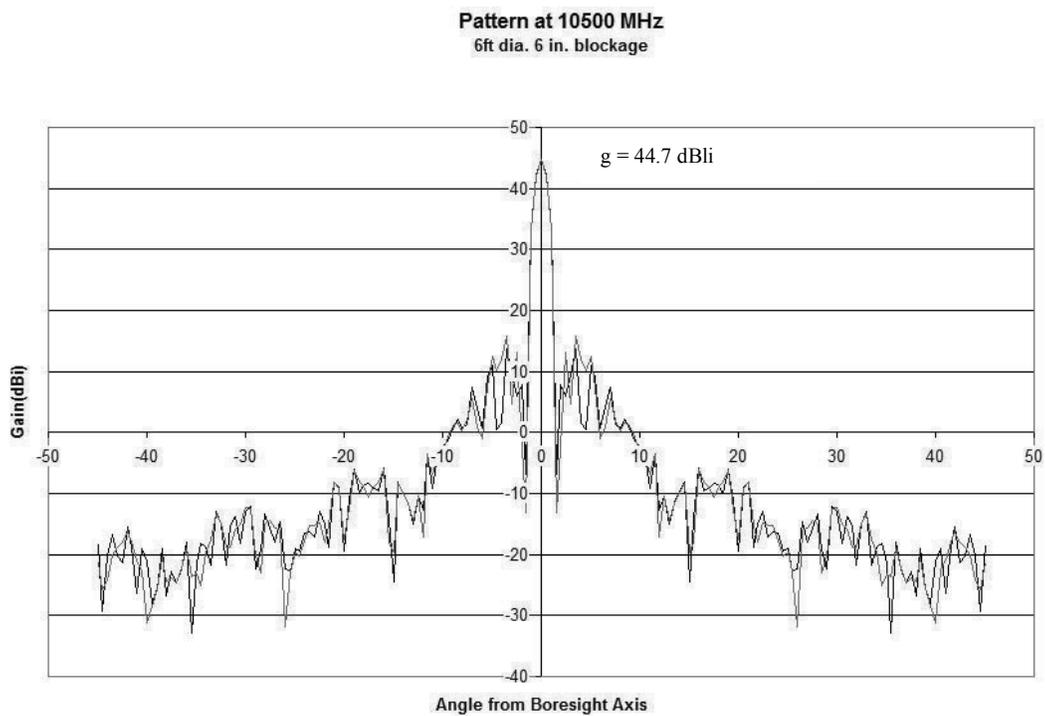
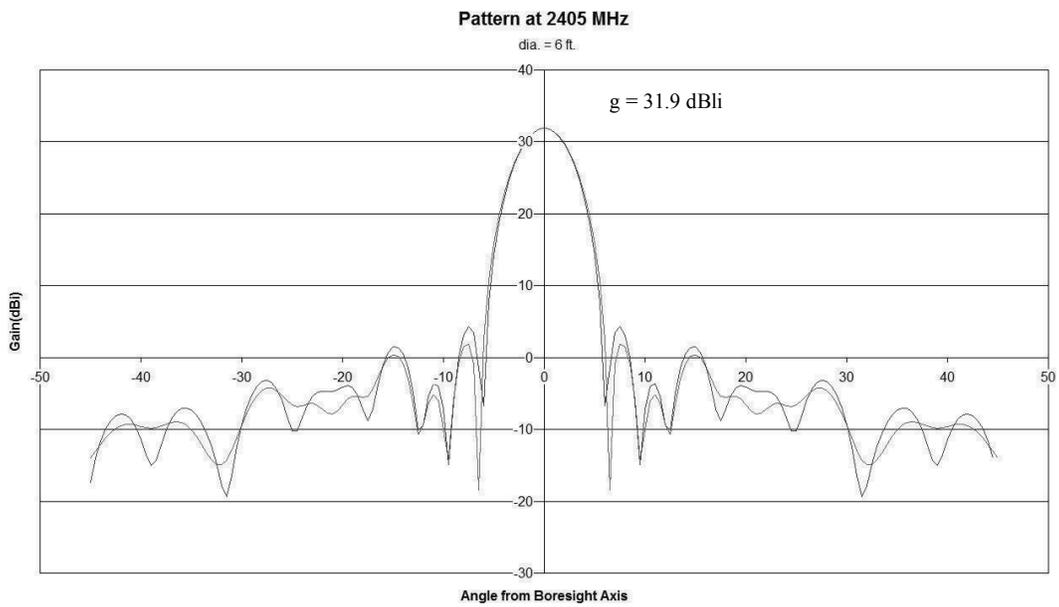


S-Band



X-Band

Secondary Patterns

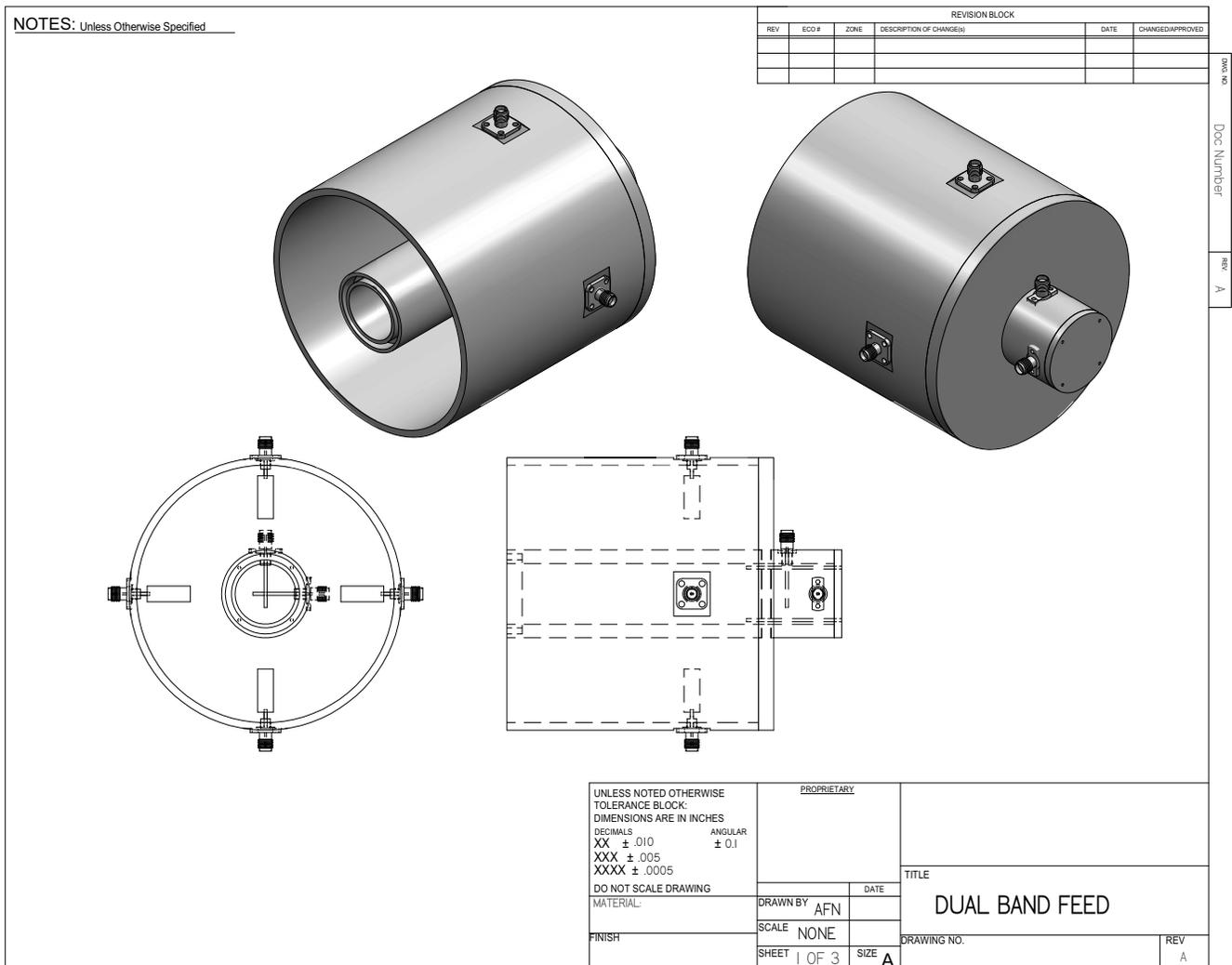


The computed gain shows an illumination efficiency of 72% for S-Band and 70% for X-Band. Sidelobes are down more than 20 dB below the main beam.

Conclusion

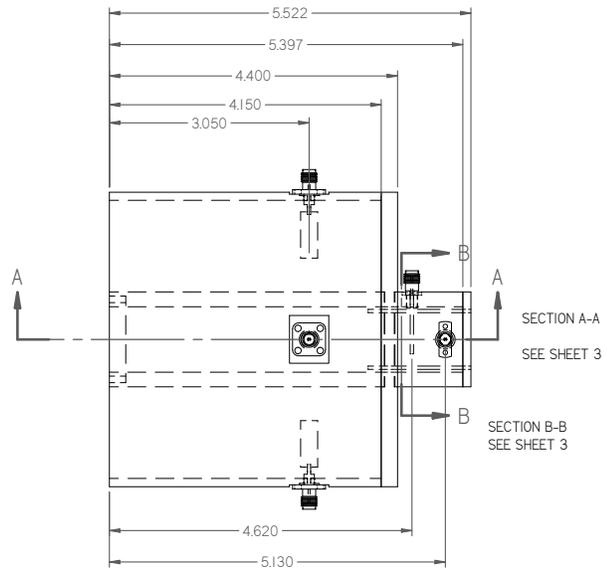
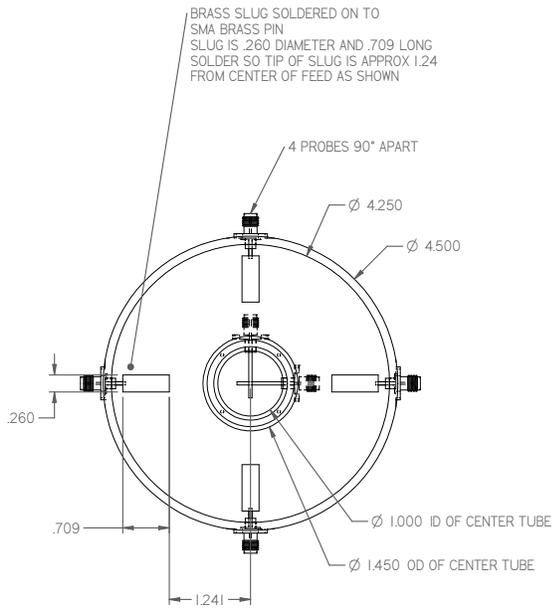
The antenna described above meets the amateur band system requirements with near text book perfect performance. Several designs of this type have been fielded and performance predicted by HFSS simulation has been repeatedly confirmed by testing.

Additional features of the coaxial feed is that it can operate simultaneously in all bands, it can be configured to operate in any set of dual simultaneous polarizations and, importantly, it remains focused at all frequencies.



NOTES: Unless Otherwise Specified

REVISION BLOCK					
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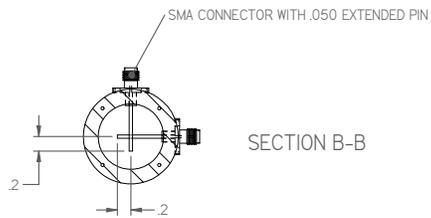
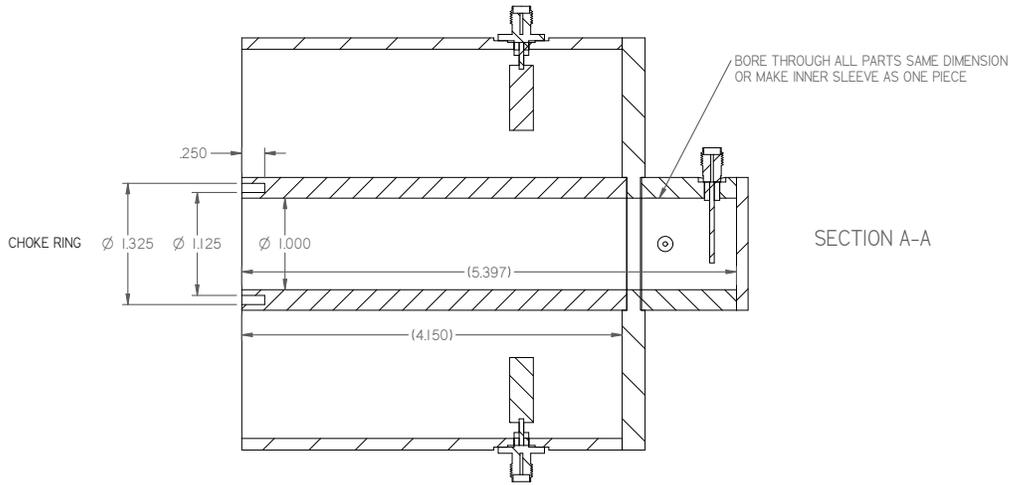
UNLESS NOTED OTHERWISE TOLERANCE BLOCK: DIMENSIONS ARE IN INCHES DECIMALS ANGULAR XX ± .010 ± 0.1 XXX ± .005 XXXX ± .0005	PROPRIETARY		TITLE DUAL BAND FEED
	DO NOT SCALE DRAWING	DRAWN BY AFN	
MATERIAL:	SCALE NONE	DRAWING NO.	
FINISH	SHEET 2 OF 3	SIZE A	REV A

DRAWING NO. REV. A

NOTES: Unless Otherwise Specified

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SCALAR FEEDS FOR PRIME FOCUS DISH ANTENNAS



Measured Results from Two Ku Band Horns Tested at 10.368 GHz

Brian Thorson, AF6NA

Member, The San Bernardino Microwave Society

Purpose

Upgrading my Amateur 10 GHz system has been an ongoing project since 2009. Having obtained a 48-inch dish antenna, I wanted to use the best feed horn I could find. Ku band horns designed for 12 GHz satellite TV reception are available from surplus sources. I performed feed pattern tests to determine if the horns I had would be suitable for my antenna. The test range, equipment, methods, data and feed horn patterns are presented in this paper.

Background on Scalar Feeds

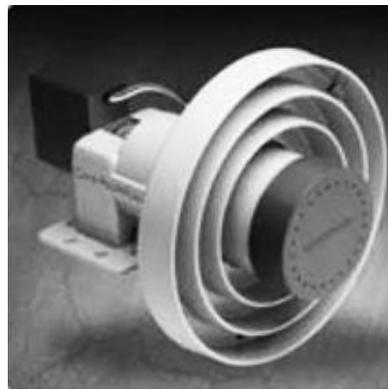
Early parabolic dish antennas used rectangular waveguides or pyramidal feed horns. These feeds inherently created unwanted side lobes resulting from edge currents at the open mouth of the horn. Affordable scalar feeds appeared on the satellite TV market in the late 70s to early 80s. The advertised features of these horns were low sidelobes and a more round, symmetrical pattern of illumination for a circular dish antenna, resulting in lower noise than conventional feed systems.



Receiving watchable TV signals was a technical challenge. The programming came from geostationary satellites at a distance of 22,300 miles above the equator. The satellite TV was down-linked to earth on the C-Band at 3.7 to 4.2 GHz microwave channels. From satellite to earth, the typical signal loss was 196 dB. Most local cable TV companies used large commercial reflector antennas on the order of 18 to 25 feet in diameter, but hobbyists and consumers were obtaining watchable TV on 12-foot or 10-foot dish antennas. The smaller consumer antennas needed high performance to pull in acceptable TV pictures. This meant minimizing the noise of the smaller dishes was important. Scalar feeds were used to accomplish this.



Prof. Taylor Howard



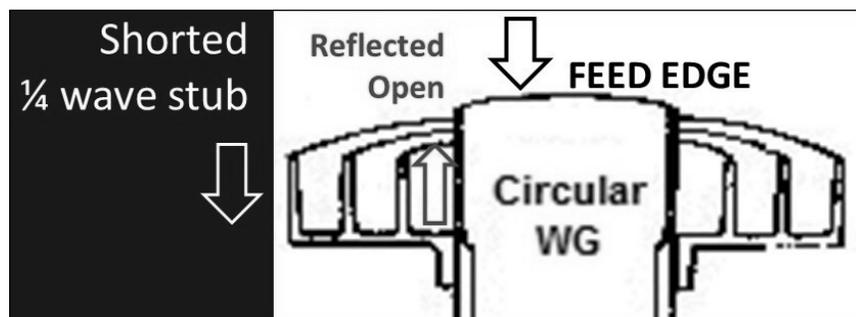
Chaparral C-Band Super Feed



**Chaparral Ku-Band
Scalar Feed**

Radio Amateurs and electronics hobbyists were among the first to design, build and operate home satellite TV receive systems for personal use and entertainment. Consumers followed in the ensuing years as home satellite TV became popular and systems were made available for sale. Electrical Engineering Professor Taylor Howard, at Stanford University, was one of the pioneers in home satellite TV and a radio Amateur (SK-W6UGL). He invented the “scalar” feed for C-Band (Above) and co-founded Chaparral Communications.

How a Scalar Feed Works



The main features of a scalar feed are side lobe suppression and a more symmetrical, circular radiation pattern. Without the rings, the radiation pattern would be oval shaped, having different radiation angles in the E-plane vs. the H-plane. So, the scalar rings re-shape the feed horn pattern, creating a circular pattern for a circular dish. Side lobe suppression is accomplished by cancelling the feed horn edge currents. As seen in the illustration above, waveguide theory tells us that a shorted 1/4-wave stub reflects an open circuit with minimum current flow at the open end of the stub. This effectively cancels the feed edge currents, reducing sidelobes.

My Driveway Microwave Feed Horn Range

I had the test gear and experience to set up a compact test range, so I decided to do radiation pattern tests at 10 GHz. The range can be seen in the following photo. Hopefully, the data could help me choose the best horn for my new system.



My Driveway Feed Horn Range

Is a 1-Meter Test Range Long Enough?

Some of us may have done gain testing of microwave reflectors that required a very long test range. The question comes up about how this kind of testing could be done on a compact range. Earlier in my

professional career, I set up and qualified a 3-meter open-area test range many times. As an EMI Test Engineer, I made measurements of RF and microwave signals for FCC and European regulations. Radiated emissions tests were required to be performed at a minimum distance of 3 meters to ensure the RF wave fronts were sufficiently flat, or “planar,” resulting in more accurate readings. ^(R3) This is an important design aspect of any RF or microwave test range.

So, in setting up my compact 10 GHz range, I had to think about what the appropriate measurement distance would be. Antenna pattern testing is usually done in the “far field” of the antenna at the frequency of interest. The larger the antenna, the longer the distance required to be in the “far field.” Far field is defined as the distance from an antenna where the radiated wave front is essentially flat. ^(R4, R5) This distance is determined by the following equation:

$$D_{ff} = 2A^2 / \lambda$$

where: D_{ff} = Distance to far field
 A = antenna aperture
 λ = wavelength

Wavelength at 10.368 GHz is 28.9 mm (1.14 inches). The largest feed horn width I tested was the 3-channel Chaparral, at 54 mm. Applying the equation above at 10.368 GHz, the far field for this horn begins at about 7.9 inches. Far field distances for the scalar horns under test are shown in the following table.

Aperture	Far Field Distance $D_{ff} = 2A^2 / \lambda$	Horn Type
20.6 mm	29 mm = 1.2 inches	³ / ₄ -inch copper pipe
44 mm	144 mm = 5.3 inches	Chaparral 2-Channel
54 mm	202 mm = 7.9 inches	Chaparral 3-Channel

Table A – Calculated Far Field Distances

With the small aperture sizes associated with feed horns and open waveguide, a 1-meter test distance puts all the horns tested in the far field. For the largest horn, a one meter range would be nearly five times the minimum far-field distance. So I set up the test range for a measurement distance of 39-3/8 inches (1 meter).

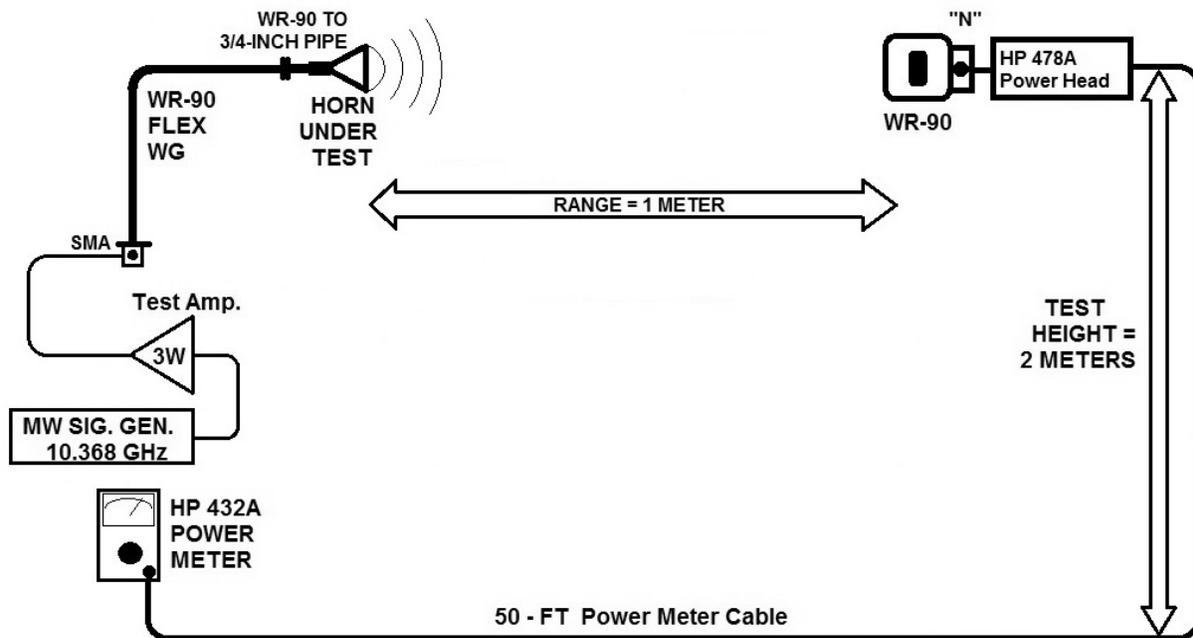
Test Equipment and Fixtures

While my test instrumentation was not “state of the art,” it was still functional and more than adequate for the testing. A list and diagram follows:

- HP 8620C Sweep Oscillator (in CW mode)
- DEMI 3WPA 3-Watt Amplifier (10.368 GHz)
- HP 5350 B Microwave Frequency Counter
- HP 432A Power Meter

- HP 478A Thermistor Power Head
- HP X281B Adapter (WR-90 / “N”)
- WR-90 Flex Waveguide (24”)
- WR-90 Right Angle Bend
- Quickset “Hercules” Tripod
- 12 V Power Supply

As can be seen in the following block diagram, the measurement system uses an analog power meter with a 50-foot power head cable. This enables power level changes to be seen immediately, faster than “real time” as defined by digitized systems.



Feed Horn Test Equipment Setup

A 3-Watt amplifier was used to compensate for the 1-meter range loss and present a measurable signal to the power meter. System dynamic measurement range is about 35 dB. While the frequency counter is not in the diagram, the generator output frequency was checked at the beginning, midpoint and after the data collection to ensure the frequency was stable within about 3 kHz.

The HP 8620 C sweep oscillator is not a “rock-solid” stable frequency instrument. It will drift in frequency by several hundred Hz. I checked this instrument with the power meter before the testing. Even though the frequency may drift a bit, the output level stays pretty constant within 10ths of a dB. It has an automatic level control circuit, so I believe it is working on this instrument.

The Horns Under Test

I have obtained two Chaparral Ku band feed horns over the past few years of collecting parts for my Amateur 10.368 GHz systems. One has two channels, the other has three. Measuring the patterns helped me decide which one to use. I also measured an open copper pipe to provide a contrasting pattern for comparison.



2-Channel Scalar



3-Channel Scalar

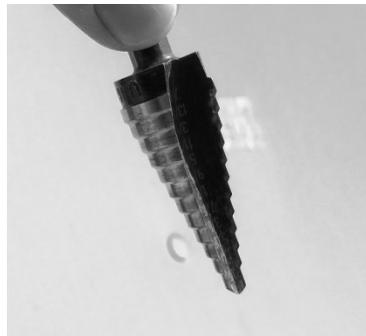


Open 3/4-Inch Copper Pipe

I expected before the testing that the two scalar horns would be pretty similar in their patterns. I include the results from an open $\frac{3}{4}$ -inch copper pipe waveguide to illustrate the advantage of the scalar horn design. Also, there were two variations on the 3-channel horn. One was drilled out to fit a $\frac{3}{4}$ -inch pipe using a combination bit. The combination bit leaves a couple of steps in the horn waveguide between the copper pipe and the horn inside diameter. I had questions about whether this double step transition affected the radiation pattern. So I had a friend machine out a second 3-channel horn for comparison. The results of this comparison are included in the data that follows.



Drilled Horn



Combination Bit



Machined Horn

The other variable in the testing was that I used tan-colored packing tape over the mouth of the horns to see if it made a difference in the patterns. Tan packing tape is what I use to keep dust and insects out of the waveguide feed system.

Test Procedure

While it may seem as though some of the steps are unnecessary, all the steps are for the purpose of measurement accuracy and repeatability. Note that the abbreviation AZ means azimuth and the rotation begins at the 10 deg. AZ position through 90 deg. AZ. Negative degree positions (-10 through -90) on the graphs that follow indicate positions beginning at 350 deg. AZ through 270 deg. AZ.

1. Zero the power meter.
2. Enable the 3W P.A. and turn on the generator output.
3. Point the tripod-mounted horn at the power meter head.
4. Adjust the tripod up and down until a peak indication is obtained on the power meter.
5. Lock down elevation position.
6. Swing the fixture right and left until a peak reading is obtained on the power meter.
7. Position the angle pointer so it points at the current position (zero degree mark).
8. Zero the power meter.
9. Adjust the generator output until the power meter indicates the 0 dB reference level.
(Any level can be used as the reference, but it should be as high as is practical)
10. Swing the tripod left (CCW) to the 10 degree AZ mark and record the meter reading.
11. Repeat for 20 and 30 degrees AZ.
12. Swing back to the center (peak) position and re-check for 0 dB.
13. Zero the power meter and adjust the generator output for 0 dB reference.
14. Swing the tripod left (CCW) to the 40 degree AZ mark and record the meter reading.
15. Repeat for 50 and 60 degrees AZ.
16. Swing back to the center (peak) position and re-check for 0 dB.
17. Zero the power meter and adjust the generator output for 0 dB reference.
18. Swing the tripod left (CCW) to the 70 degree AZ mark and record the meter reading.
19. Repeat for 80 and 90 degrees AZ.
20. Swing back to the center position and re-check for 0 dB.
21. Zero the power meter.
22. Repeat steps 9 through 19 for 9 positions to the right (CW) of the center 0 degree mark.

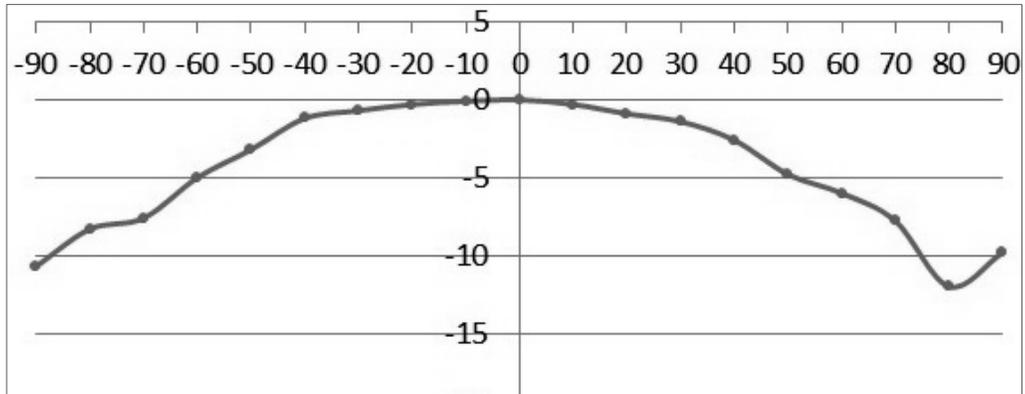
Controlling possible sources of error in radiated RF & microwave measurements is important for valid data. The two most significant potential error conditions were wind and power amplifier output level changes. When the wind was blowing, sometimes the power sensor would move, causing variations of about + and – 1 dB. In most cases when “wind wobble” was present, I simply waited out the wobble and took the data point when the power sensor was stable.

The 3-Watt power amp output power decreased when it got hot, so the temperature was controlled by removing power from the amp when measurements for one horn were completed. I also ran the (3-Watt) amp at 1.5 Watts output to reduce the heating.

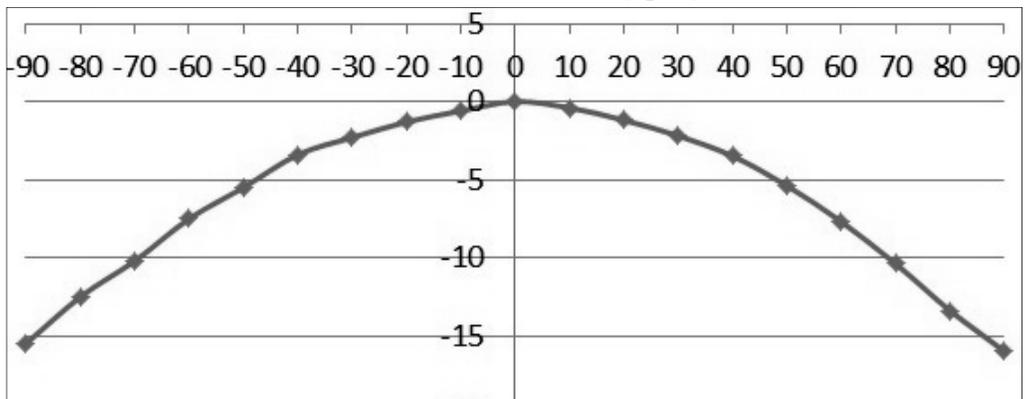
Data and Results

Data points were taken at 10 degree intervals 90 degrees right and left of the direction of the maximum power axis. My prime focus reflector has illumination angle of about 135 to 140 degrees. So, data collection between the angular values of + / - 90 degrees should include the most important values for my limited evaluation.

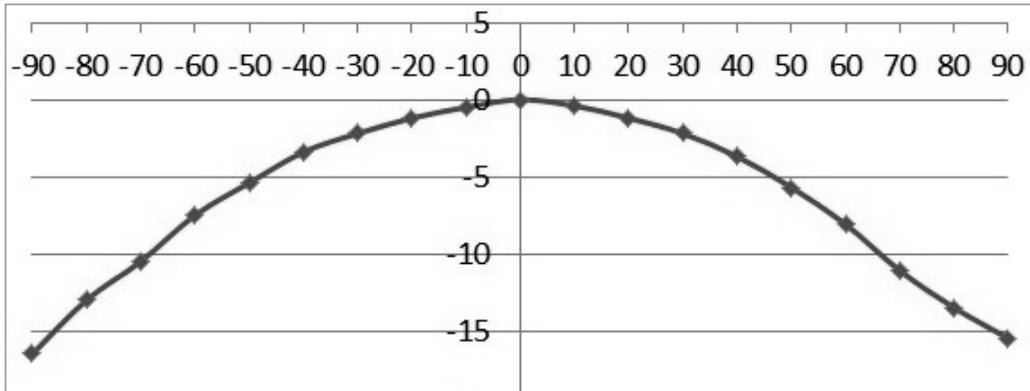
Open 3/4-inch copper pipe (open)



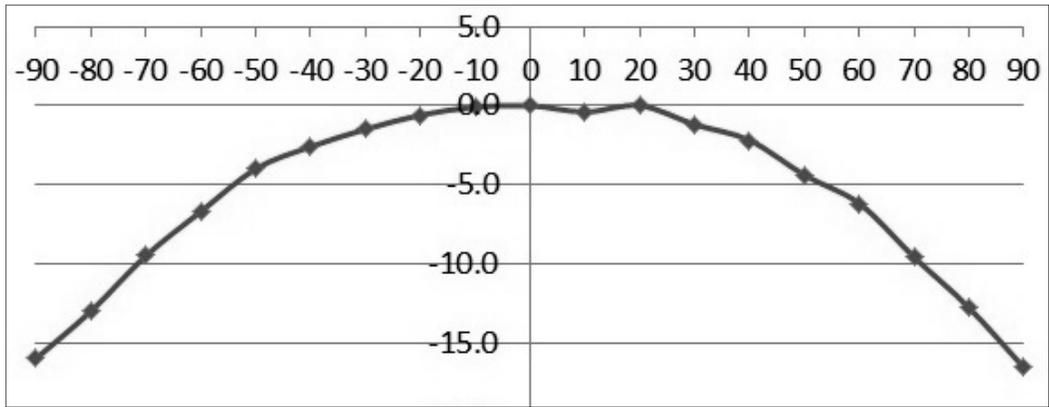
2-channel Scalar Horn (open)



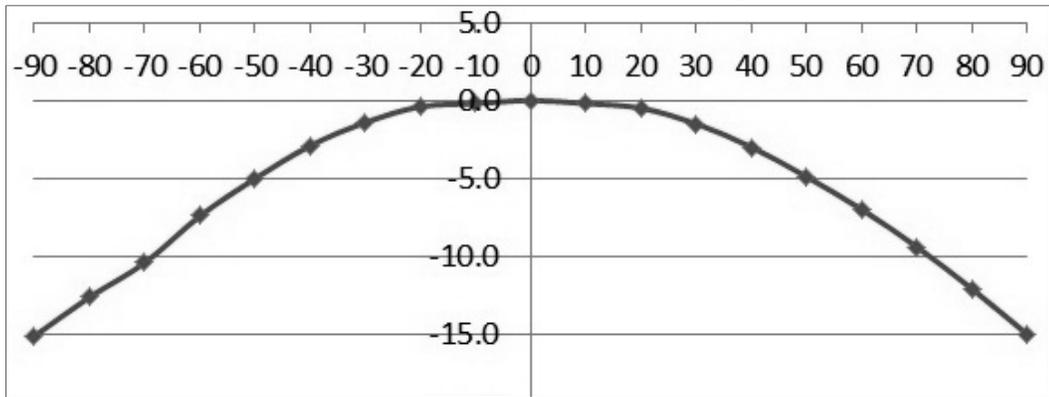
2-channel Scalar Horn (covered with tape)



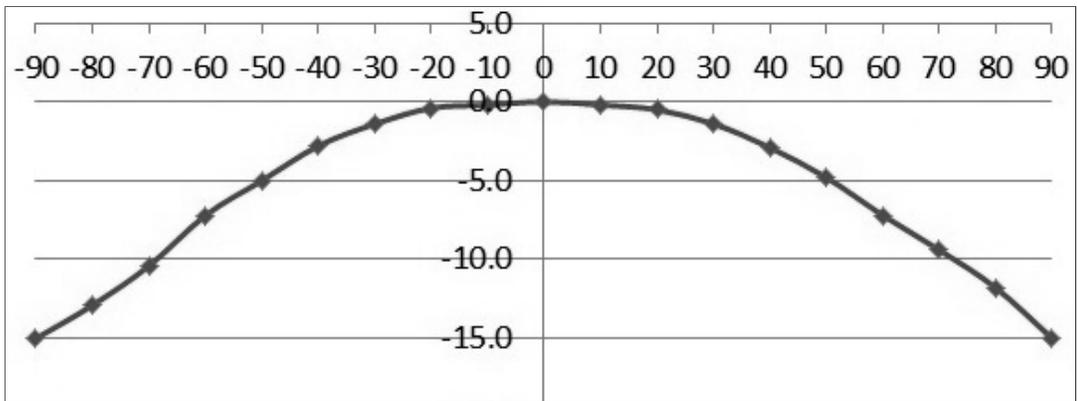
3-channel Scalar Horn (drilled – open)



3-channel Scalar Horn (machined – open)



3-channel Scalar Horn (machined – covered with tape)



Discussion / Conclusions

- The test results show that the scalar horns tested perform as they are intended to. Comparing to an open $\frac{3}{4}$ -inch copper pipe, the radiation patterns of both the 2- channel and 3-channel horns have more smooth patterns, and they show a sharper drop-off of the microwave power beyond the 130-degree illumination angle point.
- It can also be seen by close inspection of the graphs, that there is a slightly wider pattern for the 3-channel horn vs. the 2-channel. This would put more power on the reflector surface and possibly result in better efficiency and gain.
- The crudely drilled out horn seemed to have some pattern irregularities, not as smooth of a pattern as the one that was machined to fit.
- The patterns of the 2-channel feedhorn with and without tan packing tape covering the opening were very similar; too close to determine if the tape had any effect. That is good news to me, meaning I don't have to spend any significant money or time to find a plastic cap that passes microwaves freely.

Numeric data tables are presented below.

Test Data:



Horn Type: 3/4-in. copper Aperture (mm): 20.6

Flare Angle (deg.): n/a No. of Channels: n/a

<u>deg.</u>	<u>dB</u>
90	-9.8
80	-12
70	-7.8
60	-6
50	-4.8
40	-2.6
30	-1.4
20	-0.9
10	-0.3
0	0
-10	-0.1
-20	-0.3
-30	-0.7
-40	-1.2
-50	-3.2
-60	-5
-70	-7.6
-80	-8.3
-90	-10.7

Test Data:



Horn Type: 11GHz Chaparral Aperture (mm): 19
Flare Angle (deg.): 180 No. of Channels: 2

<u>deg.</u>	<u>dB</u>
90	-16
80	-13.4
70	-10.4
60	-7.7
50	-5.4
40	-3.5
30	-2.2
20	-1.2
10	-0.4
0	0
-10	-0.6
-20	-1.3
-30	-2.3
-40	-3.4
-50	-5.5
-60	-7.5
-70	-10.2
-80	-12.5
-90	-15.5

Horn Type: 11GHz Chaparral Aperture (mm): 19 /w/ tape
Flare Angle (deg.): 180 No. of Channels:2

<u>deg.</u>	<u>Db</u>
90	-15.5
80	-13.5
70	-11.1
60	-8.1
50	-5.7
40	-3.7
30	-2.2
20	-1.2

10	-0.4
0	0
-10	-0.5
-20	-1.2
-30	-2.2
-40	-3.4
-50	-5.4
-60	-7.5
-70	-10.5
-80	-13
-90	-16.5

Test Data:



Horn Type: 11GHz Chaparral – **drilled out** Aperture (mm): 19
Flare Angle (deg.): 180 No. of Channels: 3

<u>deg.</u>	<u>dB</u>
90	-16.6
80	-12.8
70	-9.6
60	-6.2
50	-4.4
40	-2.2
30	-1.2
20	0.0
10	-0.4
0	0.0
-10	-0.1
-20	-0.6
-30	-1.5
-40	-2.6
-50	-4.0
-60	-6.7
-70	-9.5
-80	-13.0
-90	-16.0

Test Data:



Horn Type: 11GHz Chaparral - **Machined** Aperture (mm): 19 w/ Tape
Flare Angle (deg.): 180 No. of Channels: 3

<u>deg.</u>	<u>dB</u>
90	-15.0
80	-12.1
70	-9.4
60	-7.0
50	-4.9
40	-3.0
30	-1.5
20	-0.5
10	-0.2
0	0.0
-10	-0.2
-20	-0.4
-30	-1.4
-40	-2.9
-50	-5.0
-60	-7.4
-70	-10.4
-80	-12.6
-90	-15.1

Horn Type: 11GHz Chaparral - **Machined** Aperture (mm): 19 w/ Tape
Flare Angle (deg.): 180 No. of Channels: 3

<u>deg.</u>	<u>dB</u>
90	-15.0
80	-11.8
70	-9.4
60	-7.2
50	-4.8
40	-2.9
30	-1.4
20	-0.5
10	-0.2

0	0.0
-10	-0.2
-20	-0.4
-30	-1.4
-40	-2.8
-50	-5.0
-60	-7.2
-70	-10.4
-80	-12.9
-90	-15.0

References:

- R1: Paul Wade W1GHZ - High-Efficiency Feedhorns for Prime-focus Dishes - VE4MA and Chaparral feeds with Septum Polarizers - ©2006 w1ghz@arrl.net

- R2: Paul Wade, W1GHZ, *Online Microwave Antenna Handbook Chapter 6 – Feeds For Parabolic Dish Antennas*, 1998

<http://www.w1ghz.org/antbook/contents.htm>

- R3: *American National Standard for Methods of Measurement of Radio Noise Emissions from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz*, ANSI C63.4, 2009

- R4: Krauss, *Antennas*, 1988

- R5: ITT, *Reference Data for Radio Engineers*, 1956

Construction Notes for Modification of “Atomic Beam” Flashlights for Optical Communications

By Barry Malowanchuk, VE4MA

At the end of May 2019 I was introduced to a novel light communications system developed by Rob K0XL and Warren WF0T. The system makes use of a simple yet excellently performing receiver developed by Clint KA7OEI and improved by Ron K7RJ. The system makes use of “page magnifier” flat lenses and 3D printed parts used to mount the components on a simple 1 in square aluminum rail. For transmit the system designed by Rob K0XL makes use of a high intensity LED flashlight. This paper details the steps I went through in making the modifications to the Atomic Beam flashlights.

The original modifications were described in brief by Rob K0XL in a paper called “PWM Optical Transmitter LED Driver” dated 07/01/2018. Diagrams are included that show the basic electrical modifications required (see Appendix). The Flashlights are available from several sources on E bay and claim very high intensity output and feature an adjustable lens that allows focusing into a tight beam or onto a lens surface. The flashlights are packaged as shown in Figure 1.



Figure 1 Flashlight Packaging

Disassembly

-After unpacking the flashlight, unscrew the bottom battery cover and the Lens assembly from the top end (see Figure 2).

-Use a “snap ring” tool to unscrew the ‘chrome colored” ring at the lens end (see figure 3), which will free up the LED assembly from the main body.

- Pry out the PCB from the back end of the LED assembly as shown in figure 4.

PCB Modifications

-Remove the 3 leaded SOT-523 surface mount part shown near the middle of the PCB in figure 4 and add solder to 2 of the now empty pads as shown in figure 5.

-Mount a plastic case 2N2222 as shown in figure 6, with the base lead but up behind it for attachment to R3 a 10 k resistor leading to the outside driver.



Figure 2 Initial Disassembly



Figure 3 Removal of LED Assembly

The resistor is mounted so that the lead passed through the small notch in the side of the board, and will be attached to a cable coming from the external driver board.

LED Changes

The LED's used in these flashlights are of the LUXEON III or "Star" type (see figure 7). The original LED produces white light. Field tests had shown a significant nuisance factor to shining white lights into the darkness and the preference is to use either or both a Red 630 nm or Infra-Red 850 nm LEDs. The IR light is significantly better as the silicon detector efficiency is better at that wavelength, but the Red LEDs allow visual verification of focusing and pointing. The LEDs are available in 1, 3 and 5 Watt versions in a variety of colors and the style used here are of the . K0XL used 3 W LEDs and the FET switching transistor seems to handle that OK. The current will increase from 600 mA to 1.2 A for the 5 W devices.

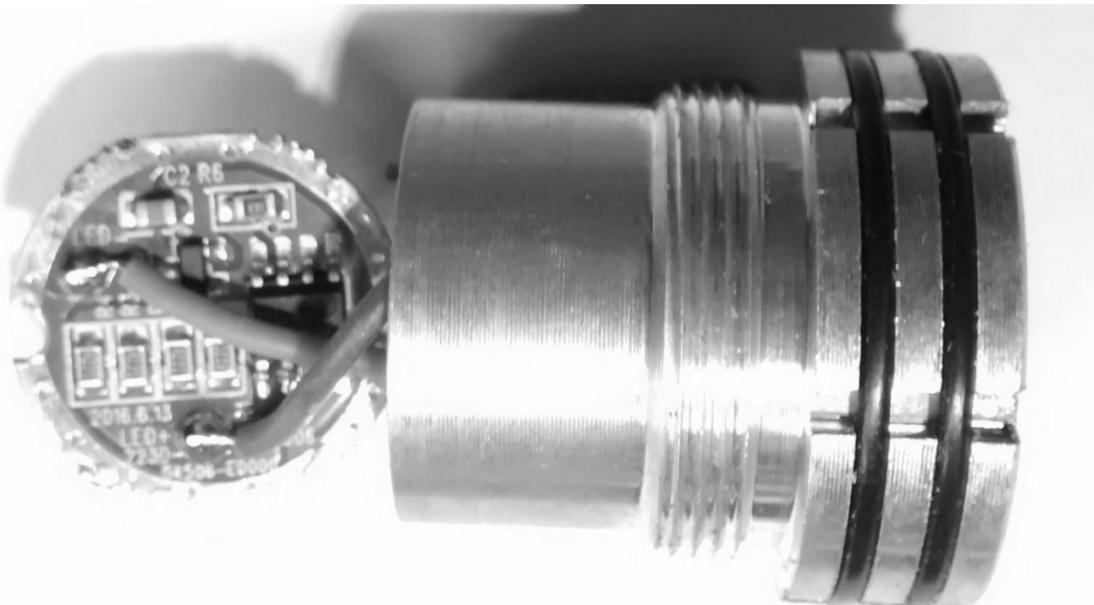


Figure 4 PCB Removed from LED Assembly

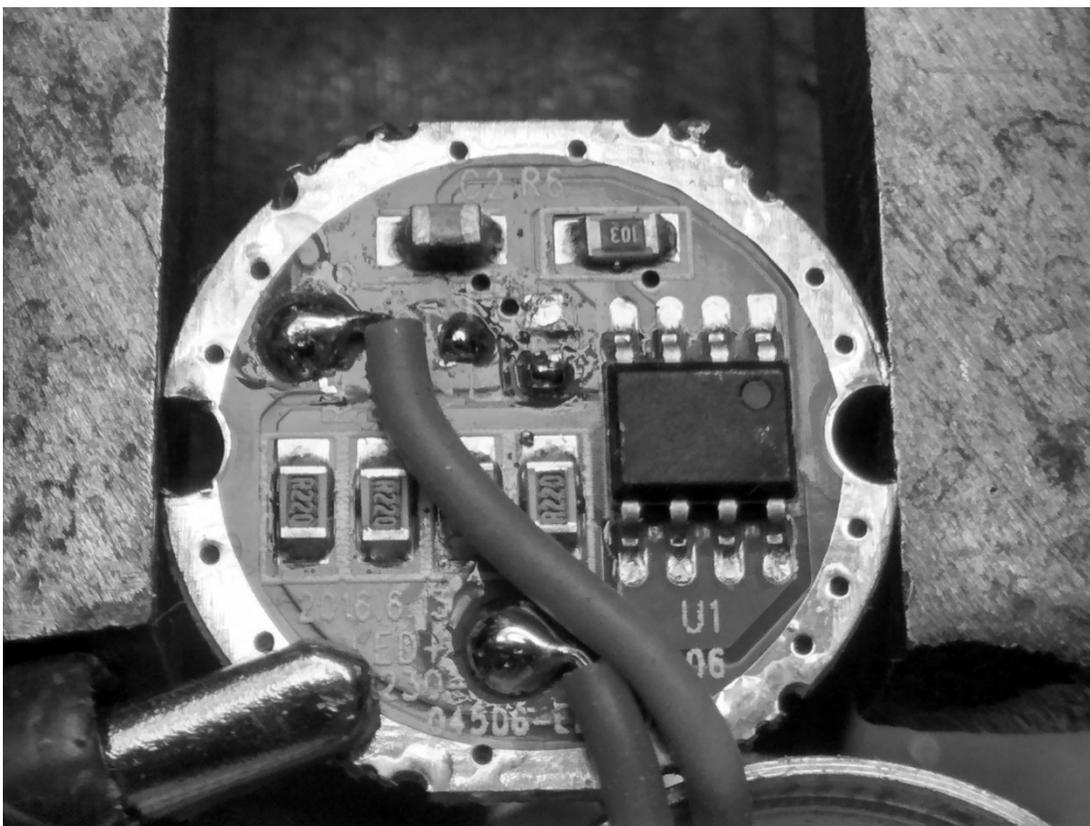


Figure 5 PCB with Surface Mount Part Removed



Figure 6 2N2222 & 10 k Resistor Added

LED Changes Continued

In order to change the LED here are the steps I used:

- Unscrew the bronze/ copper colored ring shown in figure 3 using the “snap ring” tool. You will find a clear plastic insulation disc underneath that must be reused later.
- You can now extract the LED PCB, although it may stick down with the heat sink compound on the underside.
- Remove the Red and Blue wires going to the LED and reinstall on the new LED. Add heatsink compound to the new LED, and using the clear insulation disk and the bronze/copper colored ring, secure the new LED into the assembly.

NOTE: You may need to drill out the retaining ring to avoid shorting out the diode if the insulating disk does not cover the LED contacts

- Remove the Red wire from the PCB and install a fixed resistor used to limit the LED current. I used a single 4.7 ohm 2 Watt resistor, which is marginal dissipation wise when used continuously at full duty cycle. K0XL originally used 3 smaller value resistors to accomplish this.

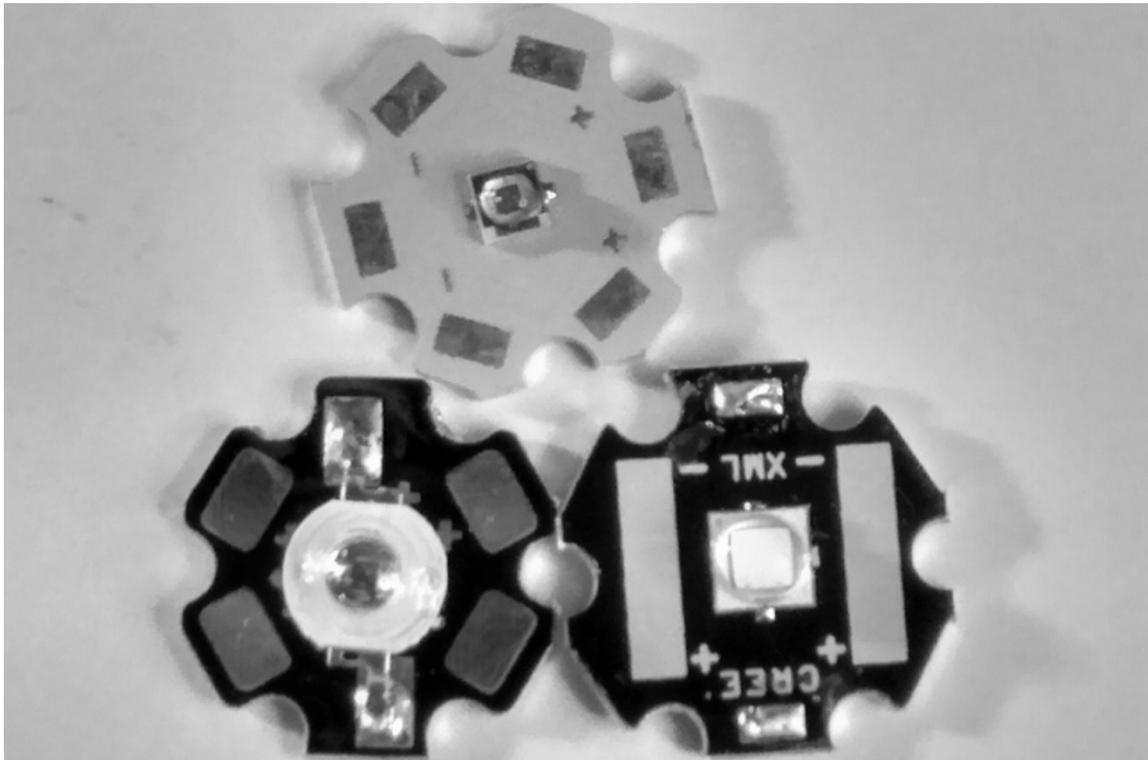


Figure 7 LUXEON III or "STAR" LEDs

Final Reassembly

The PCB can now be reinstalled in the base of the LED assembly shown in figure 4. Once the PCB is re-installed I removed the spring contact that was soldered to the bottom in figure 8. I then attached 2 wires to the bottom that will be used for the 5 Volt power supply feed.

Next a small Teflon coaxial cable was brought in to supply drive and attached to the ground and to the 10 k resistor which goes to the 2N2222 transistor as shown in figure 9.

This is a good time to test the LED assembly before Final reassembly. If all is correct the current should be approximately 600 mA with a 5 Volt supply.

Next remove the yellow button from the bottom battery cover. Now push the white push button out of the cover (friction fit) and remove entirely. Install a ½ inch grommet in the hole, which will now be used to route the 5 Volt supply and driver cables out of the final assembly.

Reinstall the LED module into the main body, tightening with the “snap ring” tool. Reinstall the Lens and retainer. Route the cables through the ½ inch grommet and reinstall the bottom battery cover to the main body.



Figure 8 Bottom of PCB showing spring to Be removed

Use of Flashlights

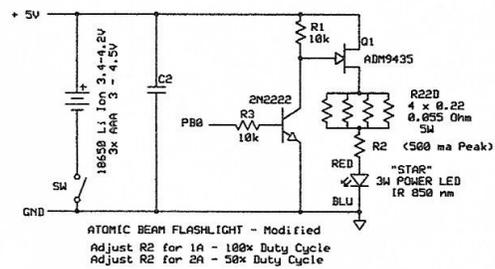
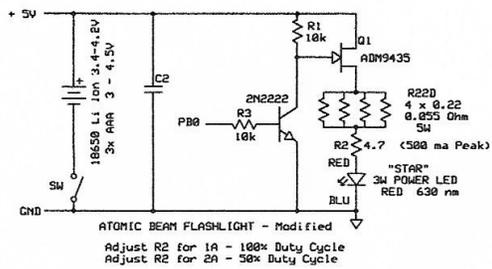
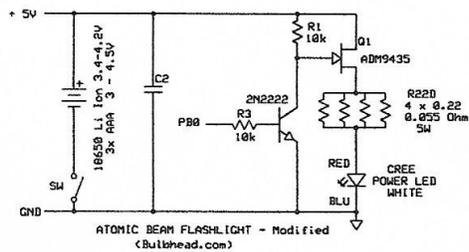
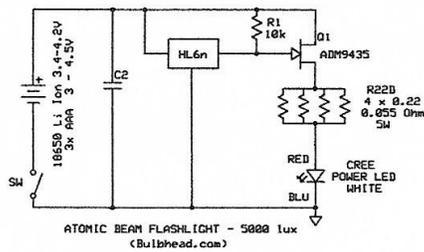
These flashlights are ideal light sources, having high output and with its adjustable lens are able to focus on one of the page magnifier Fresnel Lens in order to improve the transmit gain. I intend on using both LED types, a Red LED version initially for lining up on the path, and then changing over to an IR LED for the actual communications.

There are some higher output flashlights available but at a significant cost and it is not known how easy these will be to modify for Red or IR LEDs. This is something for future work.



Figure 9 Power and Coaxial Driver Connections to PCB

APPENDIX



Can use Internal Battery OR External 5V to power LED
REMOVE INTERNAL BATTERY WHEN USING EXTERNAL POWER
REMOVE BATTERY WHEN NOT IN USE - SW MIGHT BE ON WITH NO INDICATION

PWM Optical Transmitter
KØXL LED DRIVER Page 2
RC Jahnke v 2.0 07/01/2018

Differential GPS Azimuth Reference for Microwave Portable Operations

By David Smith VK3HZ and Rex Moncur VK7MO

We describe the use of differential GPS using carrier phase information to provide azimuth bearings for Microwave Terrestrial and EME operations to better than 0.1 degree on a 5 metre baseline. The system uses a u-blox C94-M8P evaluation kit that is available for around \$400 US. The u-blox kit is configured to provide relative positions in NED (North, East, Down) coordinates with cm-level accuracy. This position information is converted to an Azimuth bearing using an Arduino microcomputer and displayed for use as an accurate bearing reference. A rifle scope is mounted on the dish – it is aligned to the dish direction based on Sun noise. One GPS antenna is mounted directly above the rifle scope and the other is mounted remotely on a tripod. Once a bearing is determined the dish is rotated so the rifle-scope is aligned on the second GPS unit's antenna and then an azimuth ring on the dish calibrated to this bearing. Other directions can then be determined from the azimuth ring reading.

Requirement for Accurate Azimuth Bearing

The key to long distance or marginal terrestrial operation and EME with small portable stations is to remove as many variables as possible, in particular frequency and pointing. If one is waiting for a tropo-duct to open or for tropo-scatter to peak, or when using aircraft scatter which may last for only a few seconds or beaming at the moon which is not visible it is impractical to make adjustments to compensate for variables. Frequency is readily resolved by using GPS locked systems combined with Doppler correction for EME. Elevation pointing is also readily resolved with an accurate inclinometer. Azimuth can be resolved by using references such as the Moon or Sun or some identifiable feature on Google Earth such as a farm house or radio tower if available. Moderate to larger EME stations can use Moon noise but this is not effective when the Moon is close to the horizon and moon noise is swamped by ground noise. Even at high elevations Moon noise is not useful on dishes of below about 1 meter as while it can be detected it is so small that you need to be able to point accurately before you can find it. In many cases it is not possible to use the Moon or the Sun due to them being below the horizon or in cloud or to find an identifiable feature as an azimuth reference. While some claim that corrected magnetic bearings can be used the errors due to metal from cars and equipment and magnetic anomalies rule such methods out for accurate bearings. This differential GPS system has been developed to provide an almost universal method of providing accurate azimuth alignment for portable operations and remove one of the significant variables. We say "almost" because while this differential GPS system is useful in almost all situations, it does require a reasonably clear view of the sky down to 20 degrees elevation and thus it does have limitations when working in a situation such as a forest where the antennas do not have a clear view of the sky.

Required Accuracy

Antenna beam-widths are typically quoted for the 3 dB beam width. For marginal situations one would prefer to be down by no more than 1 dB. Estimated 1 dB beam-widths are set out in Fig 1 below and suggest we should aim for an alignment accuracy of better than 0.2 degrees for a 120 cm dish at 24 GHz.

Dish Size	10 GHz	24 GHz
60 cm	+/- 1.0 degrees	+/- 0.4 degrees
120 cm	+/- 0.5 degrees	+/- 0.2 degrees

Fig 1: Estimated 1 dB beam-widths

System Overview

The core of the system consists of a u-blox C94-M8P evaluation kit. The kit comprises two boards, each with a NEO-M8P-2 GPS module plus an integrated UHF link allowing the units to communicate with each other.

The NEO-M8P-2 is a high precision positioning GPS module. Traditional GPS receivers use the timing of the received signal to determine the distance to a satellite. This gives a result with typically 5 m accuracy. The NEO-M8P-2 also measures the phase of the received signal and uses this to calculate a more accurate position. The GPS signal has a wavelength of about 20 cm so there is an ambiguity regarding the exact number of wavelengths to the satellite. The 5 m uncertainty translates to 25 wavelengths. The module requires good visibility of a minimum of 6 GPS satellites to resolve this ambiguity.

With the evaluation kit, one board, referred to as the Base, is configured to firstly establish its own location accurately. It then, once per second, calculates position from the current GPS signals and sends the difference between its latest calculation and the accurately determined location to the other board over the UHF link. These are referred to as Differential Corrections.

The second board, referred to as the Rover, is configured to find its own GPS location every second and apply the Differential Corrections to the result to increase the accuracy. It then calculates the distance between two units (actually, the distance between the two GPS antennas).

Note that the location of the two boards is the reverse to what might be expected to simplify the communication of the Relative Position data from the Rover board to the Arduino system. The Rover antenna is mounted at the dish while the Base antenna is located remotely on another tripod.

The other component of the system is the Display unit. This uses an Arduino Uno with a Touch sensitive 320 x 200 pixel colour LCD shield. The Arduino is connected to the Rover board via an RS-232 receive-only connection.

Block Diagram

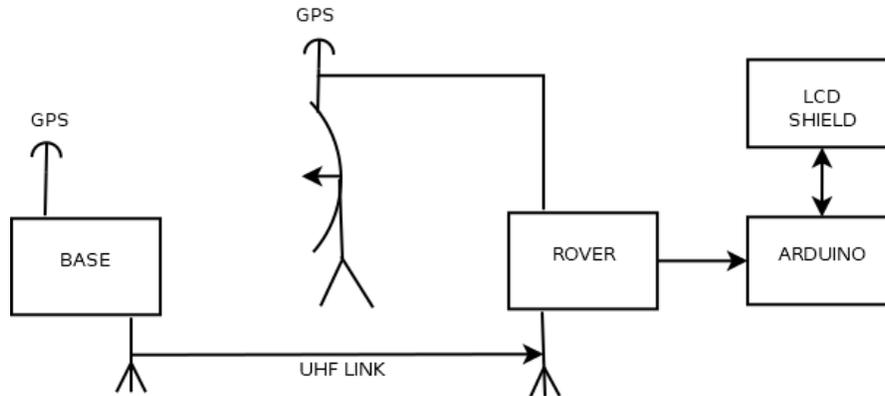


Fig 2: Block diagram of the system showing major components and flow of information.

Display

The display has two modes and one can change between them by tapping the display. The default display mode gives Azimuth, Baseline and Status as show in Figure 3.



Fig 3: This display mode shows the Azimuth (232.84 Degrees in this case), The length of the base between the two antennas (in this case 7.480 meters) and the Status (in this case 55) in brackets which indicates the bearing derives from a Carrier Phase measurement which gives the accuracy we require.

Also indicated are Status flags:

FIX: the display unit has gained a valid fix

DIF: that the display unit is receiving and using valid differential corrections from the remote unit

REL: that the relative position measurements between antennas are valid

CAR: that the bearing is based on a the relative positions determined from Carrier Phase measurements

The colour of the Status indicators changes to white when they are true.

The alternate display, shown in Figure 4, provides a graph of the Azimuth over the last 20 minutes of operation. The display auto-ranges on the vertical scale to best fit the range of the Azimuth. An average is also displayed, enabling an estimation of precise azimuth to be made.

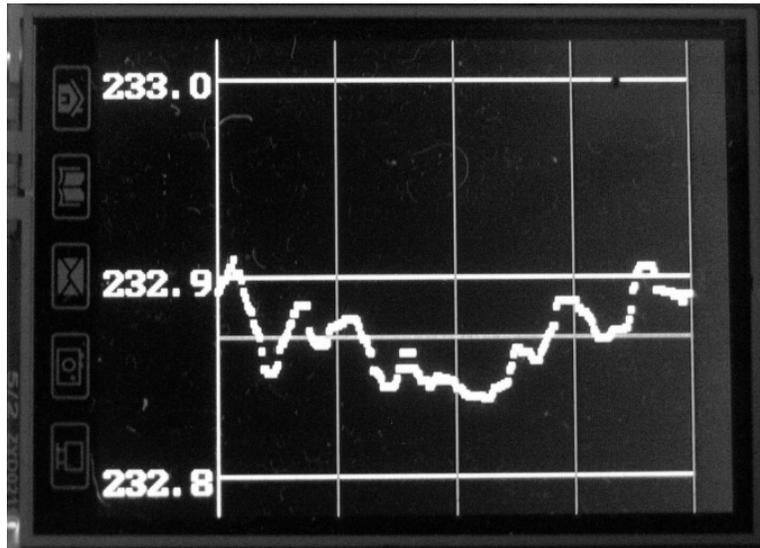


Fig 4: This is a graphical display of the indicated bearing over last 20 minutes. The green line shows the average reading which in this case is around 232.87 degrees with a variation shown by the white graph to be well within +/- 0.1 degrees.

u-blox Evaluation kit

Several variants of the kits are available to meet the frequency requirements of various countries for license-free UHF operation on either 433MHz or 915 MHz.

More information can be found at: <https://www.u-blox.com/en/product/c94-m8p>

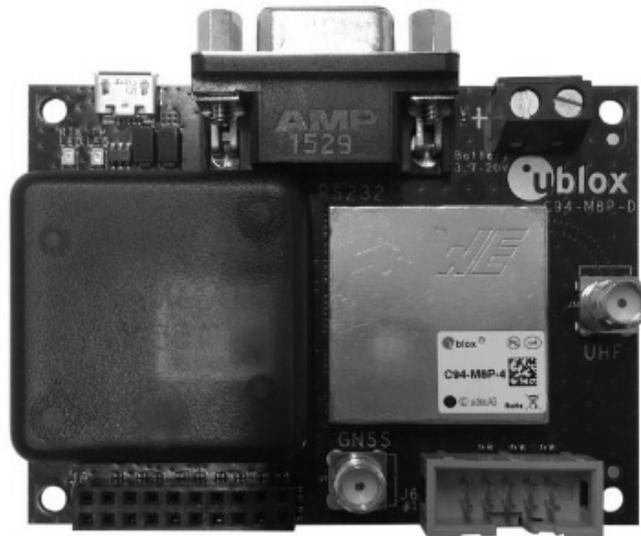


Fig 5: u-blox Evaluation Board - two identical boards come with Evaluation kit as well as cables and both GPS and UHF antennas. GPS unit under blue cover and VHF link under metal cover.

Set-up

Before the evaluation kit can be used, the boards must be configured using the program u-center which is available for download from the u-blox site. The hardware on the two boards is identical but they must be configured differently.

The first step in configuring the boards is to ensure they are running the latest firmware. Refer to the C94-M8P Application Board - Setup Guide document on the u-blox site for more details.

Next, the boards must be configured for Moving Baseline operating mode, and the messages to be sent over the communication link must be selected. Refer to the C94-M8P Application Board - User Guide document on the u-blox site for more details. Follow the steps given in section 4.3 with the following additions for the Rover unit only:

- in section 4.3.4, change Protocol Out to UBX.
- enable transmission of the UBX-NAV-RELPOSNED message as follows:
 - in u-center, Message View, select UBX then CFG then MSG
 - in the Message drop down list, find "01-3C NAV-RELPOSNED"
 - if not already ticked, click the On box next to UART1 to show a tick
 - change the 1 that appears in the box next to the tick to 5

Final thing is to save the configuration. Go to UBX CFG CFG and click Save current configuration.

The Rover antenna is mounted at the dish as shown in Figure 6. The Base antenna is located remotely on another tripod, as shown in Figure 7.



Fig 6: Rover antenna mounted on dish directly above rifle scope



Fig 7: Base antenna with u-blox evaluation board in weather proof box and separate battery

Antennas

The u-blox kit comes with two consumer grade antennas with ground planes as shown in figure 8. U-blox recommend using a better quality antenna in difficult conditions to get faster lock. While the Survey Grade antenna shown in figure 8 does give slightly improved performance, the consumer grade antennas are more than adequate.



Fig 8: The u-blox kit comes with two consumer grade antennas and a ground plane as shown on the left. On the right is a survey grade antenna which we have used for our testing.

Programming of System

The software is written in C. It runs on an Arduino Uno board.

The display is a 320 x 200 pixel colour touchscreen TFT LCD shield. Unfortunately, the display used by the authors is now obsolete. Similar displays may be found online at sites like Adafruit. Some minor changes to the software may be needed to accommodate a different display.

The software source code may be obtained by contacting David Smith VK3HZ.

Results

The graph in Figure 9 shows the results over a 5 meter baseline with a clear view of the sky above 20 degrees elevation. The graph is based on a running median value from 13 readings to reduce the impact of outliers. The peak variation is less than ± 0.1 degrees and the Standard Deviation is only 2 hundredths of a degree. It is difficult to provide an independent check of the absolute accuracy to the level that is available by this method - but as close as we can tell using Google Earth and a 15 km baseline it is within ± 0.1 degrees.

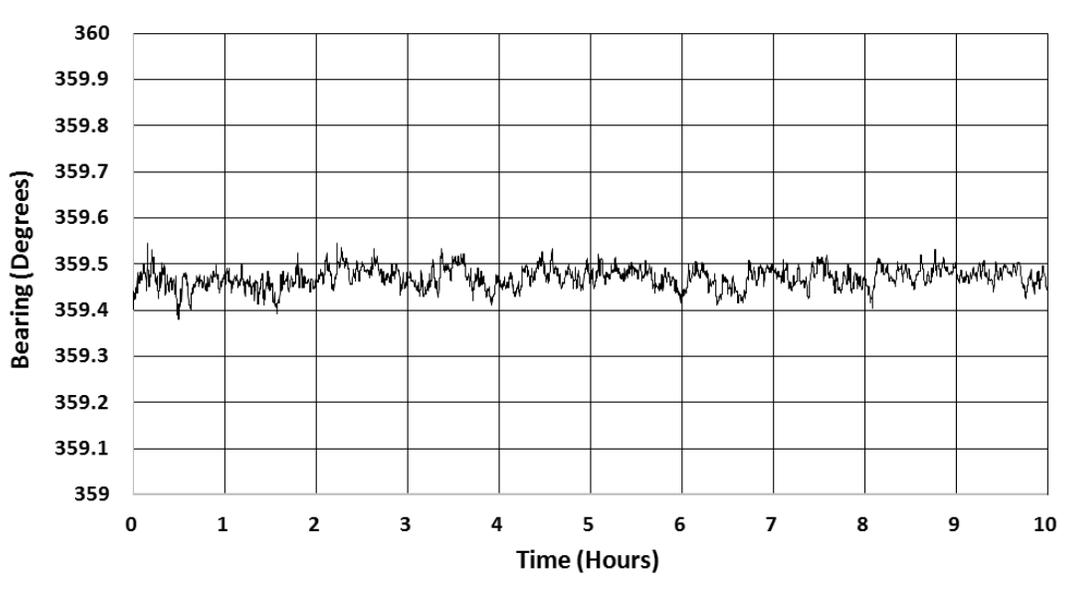


Fig 9: Results with Survey Grade antennas over a 5 metre baseline with a clear view of the sky.

In a situation where the sky is clear down to 20 degrees elevation the unit typically takes no more than a few minutes from a cold start to give an accurate bearing. The system does take somewhat longer to resolve the Carrier Phase ambiguities as the baseline is increased but works well with baselines from 5 to 15 metres. The accuracy of the bearing increases with longer baselines. We have not done sufficient testing to recommend the use of baselines beyond 15 metres, but we have no reason to believe it should not work at longer distances given time to resolve Carrier Phase ambiguities.

Impact of an obstructed sky

In situations where the sky is obstructed above about 20 degrees the system may suffer degraded performance. This occurs in two ways.

- Firstly it takes longer to gain lock and provide a bearing. For example in a situation with a clear view overhead down to around +/- 70 degrees elevation the unit did not provide a bearing at all. Under somewhat less obstruction the system may take up to 30 minutes to achieve a bearing. These impacts do vary depending on the position of the Satellites.
- Secondly, the accuracy of the bearing reduces even when it does indicate a bearing. This occurs in part because the unit tends to drop back to the 47 status message and give a bearing based on the relative position data from both units rather than the more accurate Carrier Phase as indicated by a 55 status message. Even when the unit is giving a 55 status message its accuracy can be reduced by obstruction by a factor of possibly two.

If you find yourself in a situation where you cannot gain a clear view of the sky but still have an indicated bearing we suggest you look for bearings with a 55 status message and also watch the graphical display to gain an indication of the variability.

Conclusions

Providing one has a relatively clear view of the sky this system can provide azimuth bearing references that are more than adequate for portable EME and Terrestrial microwave operations.

Ku Band Uplink MMIC PA's on 10 GHz

Alan Devlin, VK3XPD

INTRODUCTION.

For many years I have been modifying all manner of Ku Uplink hardware down to 10 GHz. The reason - the \$cost of 10 GHz GaAs Power FET's is usually prohibitive.

However, the problem with trying to modify the vast majority of Ku Band PA's down to 10 GHz lies with the GaAs FET Output Devices that Satellite Companies often use.

Ku Uplink hardware operates in the range 14.00 through to 14.50 GHz. The Manufacturers of Ku ODU usually design their Ku Band PA's with "power" GaAs FET's that have a nominal specified Operating Frequency Range of circa 13.5 GHz to 14.5 GHz.

The majority of these GaAs FET devices are IMFET's (Internally Matched FET's). These IMFET devices have been deliberately fabricated to (only) operate over a "limited" Frequency range. This has been achieved by the inclusion of Inductive and Capacitive elements within the FET structure. Semiconductor manufacturers deliberately include these "Reactive" elements to provide "stable" Gain across the nominal Pass Band and NOT beyond.

The result is that these additional Reactive components have a minimal effect within the Pass Band of the device but these same components cause the device Gain to degrade outside of this nominal Pass Band.

Our 10 GHz Amateur Band is clearly a long way outside (almost 3 GHz below) the Design Specifications of these Ku Band IMFET devices. However, myself and indeed many other Amateurs have found that many of these Ku Band IMFET's will deliver useful performance (Gain) on 10 GHz. The same scenario also applies to C Band IMFET's covering 5.9 -6.4 GHz and their reuse on 5.76 GHz.

However, there is one caveat on this Modification process! The "retuning" process around these IMFET's is often VERY tedious... requiring many hours of fiddly "Snow Flaking".

On a more positive note – I recently, I acquired a quantity of scrap 4 Watt Ku Band ODU's. Once again I had to consider the value of the Ku Band Output devices and their potential for 10 GHz Operations. And so there was that usual lingering concern about the significant effort required for "retuning" these Ku Band PA's versus the desired performance on 10 GHz.

To my great surprise, when I dismantled an ODU, I found the ODU Manufacturer had used a pair of high Gain, Phased Ku MMIC's from Eudyna. The Part Numbers used were FMM5081 and FMM5059. Oddly, the Eudyna Spec's for both these MMIC's look to be identical. They have a specified Frequency range of 13.75 GHz to 14.5 GHz, a Gain of circa 30 dB and they deliver more than 2 Watts.

With such a High Gain compared to that of a high power GaAs FET (6-7 dB), if this Ku Band MMIC could be made to work on 10 GHz it would be a good cheap alternative to that ever popular Eudyna X Band MMIC - FMM5061.

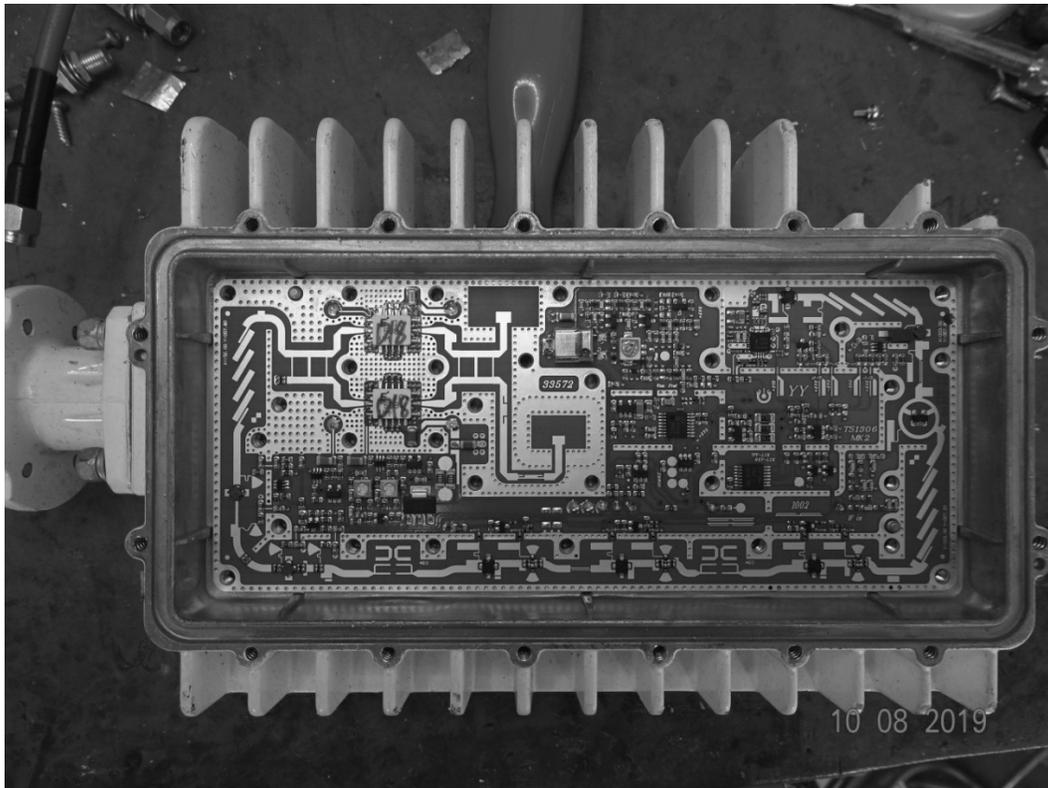
Amateur Radio Operators love the FMM5061 MMIC because of its simplicity. It delivers MORE than 2 Watts on 10368 MHz with circa 10 Milliwatts of RF drive. Although this MMIC is a very simple solution - delivering high Gain and good RF Power on 10 GHz, it has a couple of drawbacks. The unit price is fairly costly at circa \$US100 from multiple ePray Dealers and then one still needs to find/fabricate a purpose built PCB to mount it etc.

NB : Down East Microwave (DEM) used to offer this FMM5061 MMIC as a complete 3 Watt Amplifier Kit for more than \$US300 but it seems to have gone from their Sales Inventory.

TECHNICAL DISCUSSION.

With these thoughts in mind, I immediately set about modifying these Ku Band MMIC PA's down to 10 GHz.

Below is a Photo of the Ku Band ODU main board.



The 2 off Phased Eudyna MMIC's are top left. There are 4 off PCB Pads for DC Supply Rails. The two on the left are the -V_{gg} Bias Rails whilst the 2 on the right are the +V_{DD} Rails.

Note also the 2 off Wilkinson Splitters for the I/O Combining. The Input Splitter has a discrete SMD 50 Ohm Balance Load (lower left) while the Output Splitter has a coaxial "Pad" above an integrated waveguide chamber as it's Balance Load (upper right). The 14 GHz RF output is coupled via it's longer Stripline (lower right) to a coaxial "Pad" above another waveguide

chamber which ultimately couples towards the Feed Horn (Left). And finally there are 2 Tuning "stubs" on the Output Striplines immediately to the left of the Output Splitter.

All of these frequency dependent "elements" have been designed for Ku Band functionality. So... if we wish to use this MMIC in a potential 10 GHz application these "elements" must be removed (Scalpel) leaving only the I/O Striplines that feed the individual MMIC's.

MODIFICATION PROCESS.

This description has been written on the basis that any Operator can emulate the actions I have implemented on similar alternate Ku Band ODU's.

Most Ku Band ODU's operate from a single circa +20 Volt Supply. Since both MMIC's were already mounted on a PCB with Heatsinking and complete with their DC supply rails, I decided to attempt a Modification within the original case/hardware.

With the knowledge that all GaAs FET's need some form of DC isolation in their RF ports, I initially fitted SMD DC Isolation capacitors (3.9pf) in each Stripline. However, this was later found to be unnecessary because the Eudyna Datasheet shows "integrated" DC isolation was already included.

So... the next task was to fit I/O Coax tails to couple the RF In/Out of each MMIC. To achieve the best possible RF coupling, each I/O Coax needs a good Ground Plane. To achieve this I decided to solder some copper "shim" material linking as much of the existing Ground Planes as possible. While there are slight layout differences between each MMIC PCB (refer Photo below), the RF Earthing issues will be fairly obvious to see/work out.

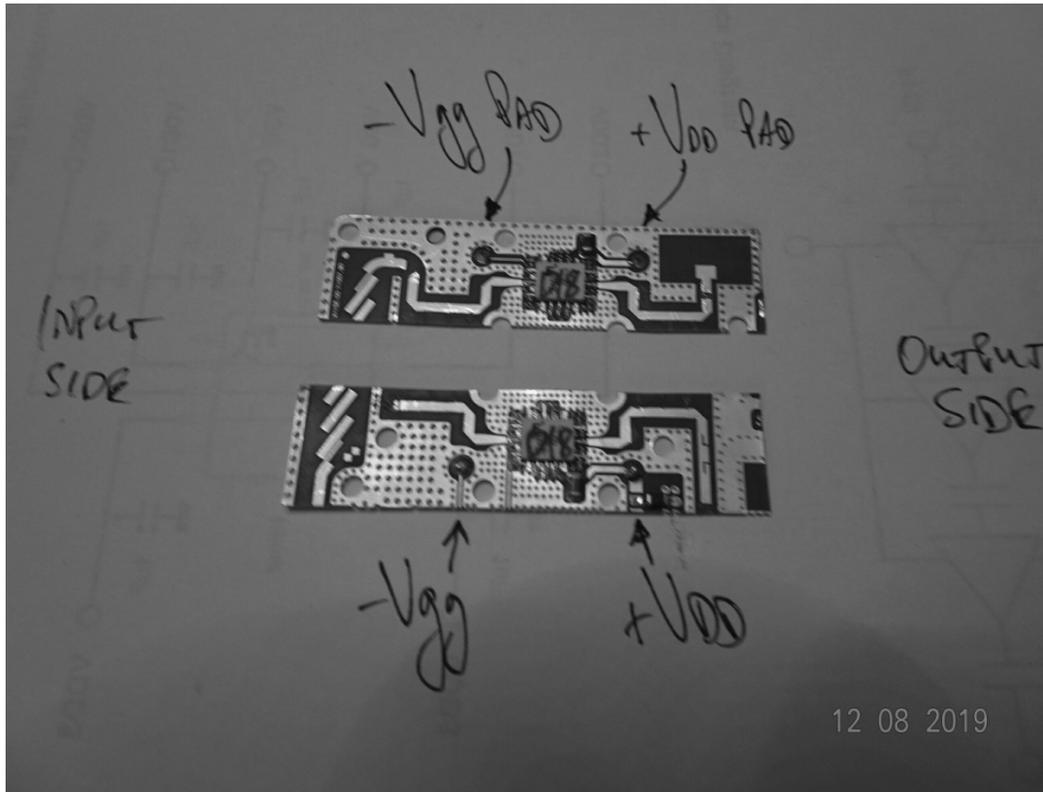
For the Coax connections, I initially tried .141 Hardline but the large physical step from the PCB Track to the centre conductor caused RF Coupling issues. The smaller .085 Hardline is a much better choice but I finally settled on using some flexible Teflon tails that I had scavenged from other equipment. I then soldered in these short Coax Tails to the I/O Strip Lines and the newly created Ground Planes.

The Power Meter (HP435A) was connected via a 30 dB Attenuator. After Powering Up the ODU, I checked the +VDD and -Vgg Supply rails and then Set the RF Drive to +13 dBm. I immediately saw a few 100 Milliwatts of RF Output. This was a very positive result for an as yet "untuned" PA. Over the next few hours I ultimately found the simplest "Snowflaking" solution that delivered the highest Gain with maximum Output.

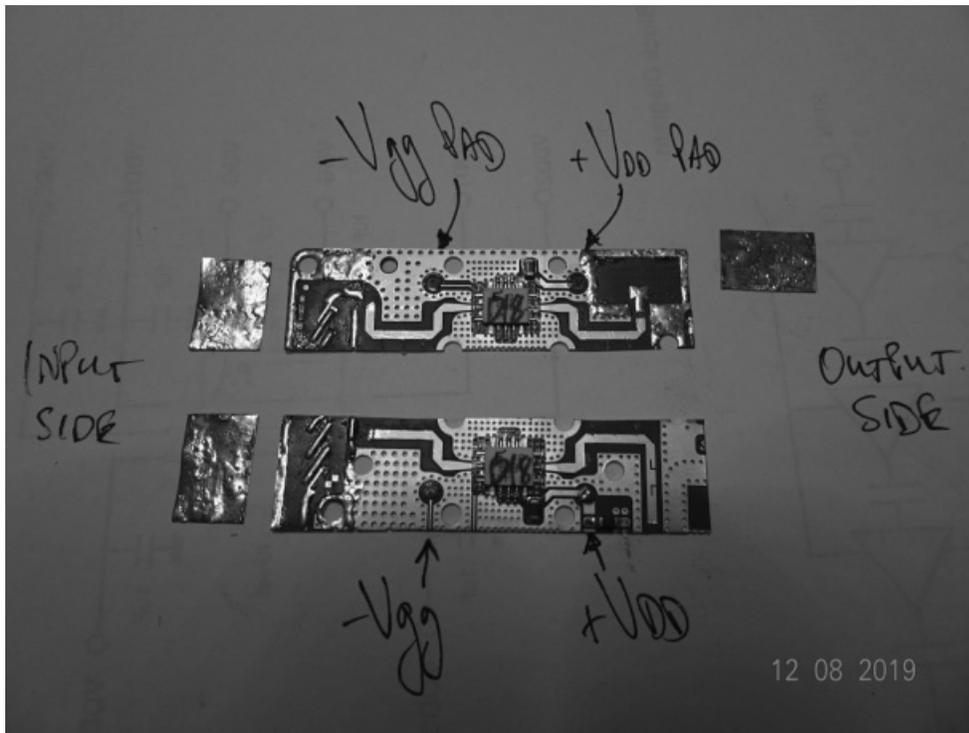
The final result was more than +30 dBm Output (1.2 Watts typical) with +13 dBm Drive. A Drive level of +10 dBm delivers circa 750 Milliwatts.

With such a pleasing result I then decided to remove (cut) these dual MMIC PA's from the Main PCB and to split them into separate amplifiers. These PCB's are VERY small !! They are only 70mm X

18mm. In USA parlance... that's less than 3 Inches long and only $\frac{3}{4}$ Inch wide. Once again, all Frequency dependent "elements" were removed leaving just the I/O Striplines feeding the MMIC's.

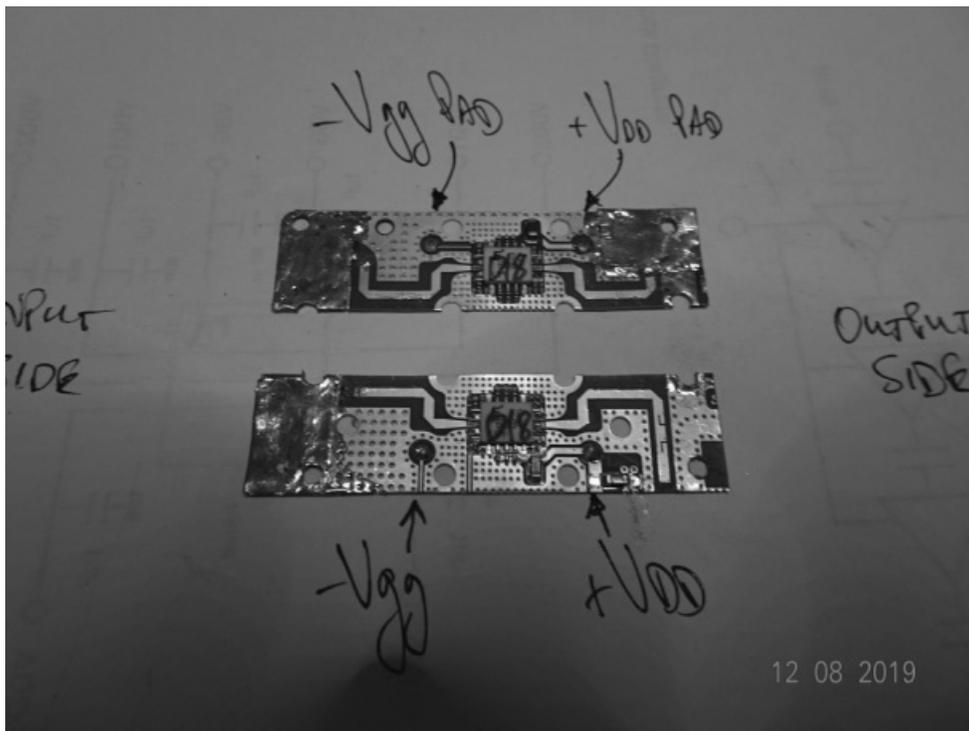


When I first mounted the top PCB to a Heatsink I discovered I had DC short circuit problems. It turned out that I had missed seeing that each PCB has Through Hole Plated DC Pads for both +VDD and the -Vgg Supply Rails. My alternate Heatsink effectively shorted these exposed Pads to Earth. To overcome this, I had to either drill a small countersunk hole in the Heatsink directly under this Pad or simply suck the solder out of this Pad and use a small Drill to eliminate the contact point of Through Hole "Pad" on the underside of the PCB before Mounting on the Heatsink. For the lower PCB, I also had to solder a very fine wire from the Vgg Bias connection to it's Bias Pad marked -Vgg. The next part of the process was the creation of the Ground Planes for each of the I/O Coax tails.



To maximise adhesion of the “tinned” copper shim material, on the Input I chose to leave all the RF Tracks intact. So... "tin" all surrounding copper tracks on the PCB and then lap the preformed shim over the entire surface and solder. Note also the I/O Striplines have been trimmed ready for their RF connections.

Below is a Photo of the completed Ground Planes.



Readers will hopefully note that I have restored (created) the PCB Mounting holes. I suggest 8 holes are needed for best RF Grounding. I cannot emphasise enough how important it is to fit all 8 Screws (3mm ?) or Self Tappers to ensure best RF Grounding. Poor RF Grounding means reduced Output !!!

A mandatory smear of Heat Sink Compound under the MMIC is also necessary to achieve optimum heat transfer.

After carefully mounting the PCB on a suitable Heatsink, I then fitted the I/O Coax tails and repeated the "Snow Flake" (retune) task. For the most part, there are only 3 or 4 Snowflakes needed to optimise this PA.

The final result can be seen here in this Photo of a single Amplifier.



This Amplifier requires 2 Supply Rails... a Negative Bias Rail V_{gg} (Green wire) of circa -0.20 Volts and +VDD Rail in the range +5 to +6 Volts maximum.

The +VDD Rail is simply a short piece of hookup wire (Orange wire) on the O/P side of the MMIC. A similar arrangement can be adopted for the V_{gg} Supply. Note that for this PA the -V_{gg} Bias Rail (now) appears on an SMD capacitor that I soldered directly to the PCB.

To achieve the quite low -V_{gg} Voltage Rail of circa -0.20 Volts, I used a -5 Volt Supply developed from a 7660 Chip or similar. This -5 Volts is coupled to a 3.9K Resistor in series with a 1K Pot. The centre of this Pot delivers a maximum of -1.0 Volts. The reason for this Bias arrangement is because the Absolute Maximum Bias voltage for this MMIC is -3.0 Volts.

Before I connected the +6 Volt VDD Rail, I preset the -V_{gg} Rail to circa -1.0 Volts to “Set” the PA at Cut Off. This ensures the MMIC will not destroy itself if the +6 Volt VDD Supply rail is NOT Current limited (i.e. 2 Amps max).

The recommended quiescent IDD for these MMIC's is 1200 mA. To achieve this IDD, I carefully reduced the -V_{gg} voltage whilst monitoring the +6 Volt IDD current. The final -V_{gg} Bias voltage was circa -0.20 Volts. The application of RF Drive will push this IDD current up towards 1.5 Amps and more, hence the need for an adequate Heatsink or small Fan.

MMIC PA AVAILABILITY.

These modified MMIC PA's are a VERY simple way to get good RF Power on 10 GHz.

To inspire more interest in 10 GHz - I am offering unmodified PCB's (only) for FREE. All I ask is for the Operator to cover the Postage costs. Enthusiastic Operators with the desire to build their own gear will then need to create the Ground Planes, fit the I/O Coax tails, mount the PCB on a suitable Heatsink, tune the Striplines (Snow Flaking) and fabricate the +6 Volt VDD and -5 Volt V_{gg} Supply rails. Aside from this Article - I can/will supply a brief Fact Sheet on how to Snowflake (location) these PCB's.

For those Op's who don't wish to Modify a PCB or perhaps do not have the Test Gear or are not confident with the Snow Flaking process, I have already modified more than a dozen of these PCB's (only).

Operators will then only need to carefully Mount the modified PCB on a suitable Heatsink etc. These complete tested 1 Watt, 10 GHz PA's (PCB's only) sell for \$Au50 (\$US35) plus a bit for P&P. I will guarantee every modified PA with a Replacement or Refund if a/any unit fails to meet expectations.

Please be aware that the addition of the Ground Planes and the I/O Coax tails has resulted in these modified PA's not looking the tidiest Amplifiers around. But... who cares !!!!

My attitude is this - so long as these PA's work as promised... then how they "look" is not a major issue. Of greatest importance to me is/was my goal of re-purposing this scrapped RF hardware.

These cute little Amps are definitely a VERY simple and cheap solution for a high Gain 10 GHz Amplifier that delivers more than 1 Watt with a modest RF Drive level up to +13 dBm.

If these (10 GHz) MMIC PA's interest you.... drop me an Email to alandevlin@bigpond.com

Cheers,

Alan - VK3XPD

Timing for Portable Digital Operation

By Rex Moncur, VK7MO, & Larry Hower, VK7WLH

We describe a simple method of providing GPS locked time to better than 0.1 second accuracy based on using an inexpensive USB GPS receiver. A warning - some USB GPS receivers do not reliably produce UTC time and we recommend the Ublox r6010 or r7020 GPS receivers as packaged by Gmouse, which our testing has shown are reliable.

Issues for Portable Operation

When operating portable the internet is often not available and may not be reliable. Thus some independent and accurate source of time is required. In the past we have used a Garmin GPS 18 with a 1 pulse per second (1PPS) output which interfaces through a serial port with a program called NMEATime. This system works well with the XP operating system. Today's PC's rarely have a serial port and even when they do we have been unable to get this system to work on Windows 7 or 10. The solution is to use a new program called NMEATime2 and a USB GPS receiver as discussed later in this note.

What timing accuracy do we need?

The WSJT slow modes are offset towards the EME delay and have a much lower timing tolerance when used for terrestrial work:

- JT65 -3 to +5 seconds
- JT4f -1.4 to +6 seconds
- QRA -1 to +4 seconds

For example if using QRA for terrestrial an error 1 second can result in one way decoding. In the worst case an error of 0.5 seconds at both ends will cause one way decoding. When looking for a weak signal to rise out of the noise it is useful to have accurate timing as good syncs based on the correct DT and DF are a useful indication of the presence of a signal. When using averaging on JT4 the timing stability needs to be better than 0.2 seconds to be accepted by the averaging algorithm. Overall we consider one should aim for a timing accuracy of 0.1 seconds.

Testing for Time Errors

To test timing accuracy we send JT messages between two computers one of which has accurate time, such as can be provided by an XP computer with 1 PPS or from a known accurate internet source. We use the WSPR mode as it averages the DT over two minute periods to improve the accuracy. K1JT has added another decimal place to the ALL_WSPR file so it reads to 0.01 seconds. However, the DT resolution of WSJT-X is limited to 0.04 seconds.

A Solution for Windows 7 or 10 using a USB Port

The author of NMEATime has produced a new version NMEATime2 that is designed to work with cheap USB GPS receivers with a USB output. It can be downloaded at:

<http://www.visualgps.net/#nmeatime2-content>

The program is available free for 30 days use and there-after requires a one-off registration fee of around US\$ 20.

NMEATime2 works by detecting first letter of the NMEA sentence and uses a long time constant loop to remove jitter. It has a claimed accuracy of 5 ms after calibration .

The NMEATime2 web site lists a number of GPS receivers that are claimed to have been used successfully. We initially chose the BU-353-S4 - USGlobalsat from this list but found it does not reliably give accurate UTC time.

GPS and UTC time

GPS units work on GPS time which is currently 18 seconds out to UTC time. In our case we need accurate UTC time. This difference increases by 1 second roughly every 19 Months, but is corrected when necessary on 30 June or 31 December by international agreement. Cheap GPS receivers set the initial GPS to UTC time difference in what is called an almanac at the time of manufacture. For example the GlobalSat unit was initially two seconds out up to 31 December 2016 and then 3 seconds out. This error is normally corrected when the GPS unit downloads what is called an ephemeris which should arrive within 12 minutes of switch-on. Accordingly, we would expect that if we wait 12 minutes before operating the UTC time should be correct. However, extensive testing of the GlobalSat unit shows that it does not reliably make this correction and about 25% of the time it is switched on it can remain 3 seconds out.

Gmouse with Ublox r6010 or r7020 GPS

In discussions with the author of NMEATime2 he suggested we try the Ublox GPS receivers to overcome the timing reliability problem. Extensive testing has shown that the Ublox GPS receivers reliably determine UTC time given 12 minutes after switch-on to gain the ephemeris and the accuracy is within the limits of the WSPR resolution of 0.04 seconds with calibration.

Calibration

NMEATime2 has a calibration facility to allow for the time delay to the first NMEA letter. For calibration one needs an accurate time reference such an internet time server of known accuracy. We have found that the calibration correction is generally no more than 0.05 seconds and the accuracy is still within the 0.1 second goal even without calibration.

Drivers for Ublox

Drivers are included in Windows 10 but need to be downloaded from Ublox site for Windows 7 at the URL below:

[https://www.u-blox.com/en/product-resources?f\[0\]=property_file_product_filter%3A2779](https://www.u-blox.com/en/product-resources?f[0]=property_file_product_filter%3A2779)

Conclusions

- Use NMEATime2 with a Gmouse unit with uBlox r6080 or r7020 and you should be within 0.1 seconds after 12 minutes, without calibration.
- Accuracy can be improved with calibration.
- This gives a simple and reliable system that works with and is powered by a USB port.

Audible Field Strength Monitoring Enables Remote Measurements

Sam Green, WØPCE

w0pce@arrl.net

I find it useful to build and use test equipment that presents measurement data to the user as a variable audible pitch (frequency) in place of, or in addition to, a visual display. Analog meters and digital displays that present data for visual observation require the dedicated use of two valuable resources: your vision and your attention. I offer an alternative method that lets you divert your vision and much of your attention to another task, such as holding test probes in place and adjusting a critical circuit parameter, while letting you hear the result. The human ear easily differentiates very small frequency changes, allowing very high analog resolution and thus high sensitivity to small changes.

One recent example uses an ATtiny85 CPU as an accessory added to a digital multimeter to enable precision analog resolution^{i,ii} from the digital instrument. Earlier, voltage-to-frequency converters such as the LM331 were the basis for such applications.

The instrument described here is an RF field strength meter with the output as an audible pitch based on the LM331 voltage-to-frequency converter. I meant this to be an instrument for the amateur radio community to make remote measurements of relative field strength. I thought it would prove useful to monitor radiated signal level at some distance from a transmitting source for antenna comparisons or other far field measurements. The output is in the 300–3000 Hz audio range for transmission back to the listener. I built versions with an AD8307 logarithmic detector that works up to 500 MHz and with an AD8313 that works up to 2.5 GHz. I also suggest a version based on the AD5513 from a previous articleⁱⁱⁱ that would work up to 4 GHz. **Figure 1** shows the circuit built into a pair of Altoids tins with the circuitry in one compartment and the power source in the other. This version used 4xAAA cells and a low dropout 5 volt regulator, but the version with 3xAAA cells and no regulator in the schematic below is simpler and works fine.



Figure 1 Circuitry and batteries built into a pair of Altoids tins

I walked around my workplace with the AD8307 prototype and listened to different sources of RF. When I heard the data exchange as an RF reader interrogated my employee badge, I got the idea that this could become a cell phone and Wi-Fi detector. That might prove useful to sense such devices. I built up another prototype using an AD8313 logarithmic detector with 2500 MHz response to sense cell phones and Wi-Fi. With it I could tell which laptop had its wireless enabled when I walked up and down the aisle on a commercial airline flight with headphones connecting to the unit in my pocket. Because airlines want RF emitters turned off, and because I worked for an airplane company that offers incentives for intellectual property, I patented this method. See the details in patent 7,898,395 online. Boeing owns the patent, but hobbyists can use the concept to build their own.

In the shack, you read the meter of an ordinary instrument, but remote operation requires a way to return measurement data to the user while making adjustments that affect the remote reading. Conversion of the field strength reading to a varying audible pitch allows return of the measurement data via an audio modulated radio link or a telephone line (I started using this technique in the late 1980s to monitor the photocurrent of a remote optical receiver in a laser communications link with the microphone end of a telephone handset in a modem speaker cup!).

How it works

Figure 2 shows a block diagram of the audible field strength monitoring instrument.

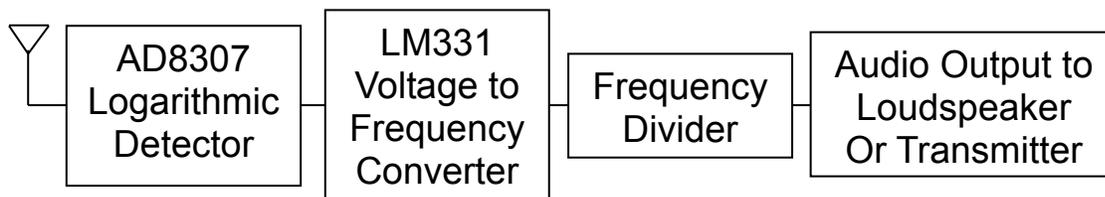


Figure 2 Functional Block Diagram of RF Field Strength Meter with Audible Output

The Analog Devices AD8307 accepts a 90 dB input dynamic range of RF signals in the audio to 500 MHz range and compresses it into a monotonic DC voltage output range with about a 2 Volt excursion. I selected parameters to match the output voltage range to obtain the desired 300 to 3000 Hz audio frequency range from the voltage-to-frequency converter that matches the audio band-pass of both the telephone system and amateur radio FM transceivers.

Voltage-to-frequency converters output narrow pulses with low audible energy. I set the frequency twice as high and divide the output by two with a toggle flip-flop to yield a rectangular wave with high audible energy. A type D flip-flop with the not Q output connected to the D input forms a toggle flip-flop. Two flip-flops in parallel offer twice the drive current to the built in speaker. **Figure 3** shows a semi-pictorial schematic of the 500 MHz instrument as a bottom view wiring guide. Adding a 1 μ F capacitor from OFS pin 3 to ground enables the AD8307 logarithmic detector to work down to low audio frequencies as well.

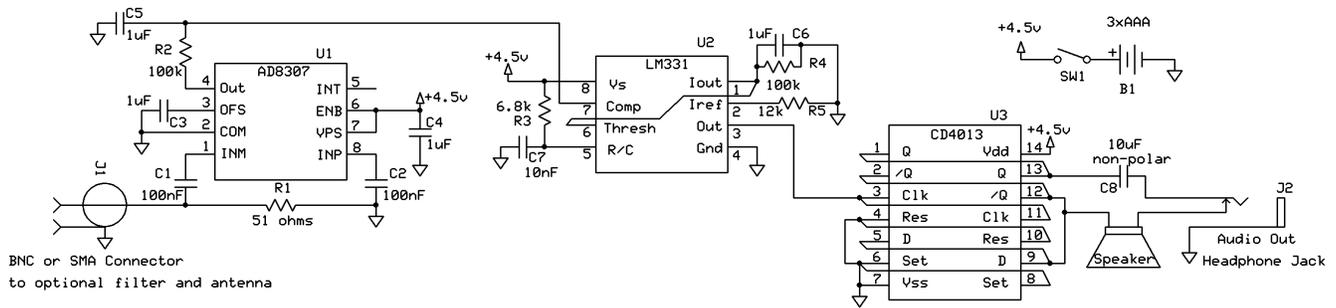


Figure 3 500 MHz version of Audible Field Strength Meter using AD8307

Figure 4 illustrates example circuit waveforms at the wide dynamic range RF logarithmic detector input, the output of the logarithmic detector to the voltage-to-frequency converter, the output pulses from the voltage-to-frequency converter, and the output of the toggle flip-flop frequency divider. The logarithmic RF detector accommodates an immense dynamic range, making this instrument much more useful than one based on a simple diode detector.

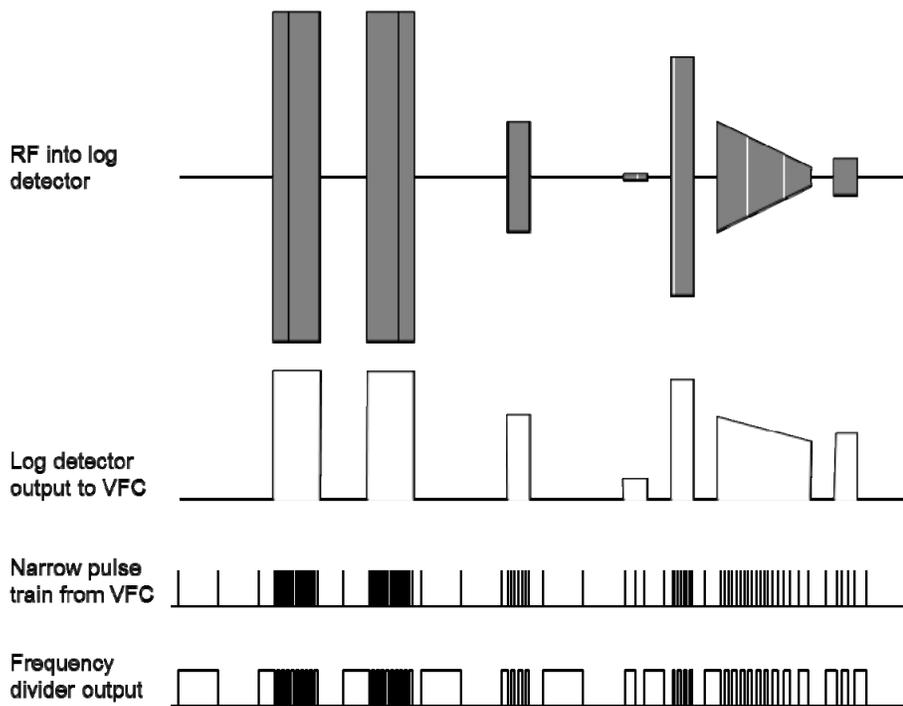


Figure 4 Example Circuit Waveforms

Microwave Versions

Figure 5 shows the front-end for the 2.5 GHz version using the AD8313 logarithmic detector.

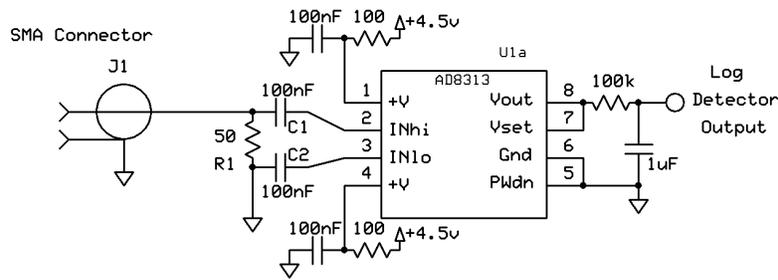


Figure 5 Front-end for Audible Field Strength Meter using 2.5 GHz AD8313

Figure 6 shows the front-end for the 4 GHz version using the ADL5513 logarithmic detector.

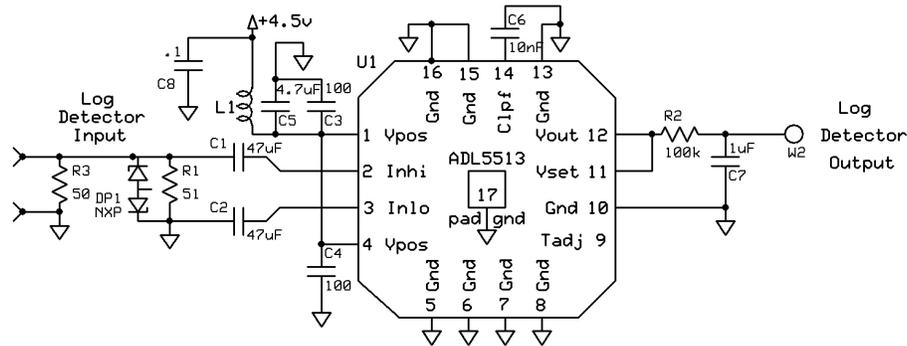


Figure 6 Front-end for Audible Field Strength Meter using 4 GHz ADL5513

The microwave versions hear the digital transactions of my Wi-Fi router. These units prove very useful when you want to hear whether an RF source is operating. I use them in my work frequently and think you will find them useful.

Dr. Sam Green, WØPCE, is a retired aerospace engineer living in Saint Louis, Missouri. He holds undergraduate and graduate degrees in Electronic Engineering from Northwestern University and the University of Illinois at Urbana respectively. Sam specialized in free space optical and fiber optical data communications and photonics. Sam is currently designing and prototyping innovative targets for guns with laser bullets. Sam became KN9KEQ and K9KEQ in 1957, while a high school freshman in Skokie, Illinois, where he was a Skokie Six Meter Indian. Sam held a Technician class license for 36 years before finally upgrading to Amateur Extra class in 1993. He is a member of ARRL, a member of the Boeing Employees Amateur Radio Society (BEARS), a member of the Saint Louis QRP Society (SLQS), and a member of the Bi-State Amateur Radio Society. Sam is a Registered Professional Engineer in Missouri and a Life Senior Member of IEEE. Sam is named as inventor on eighteen patents.

ⁱ Add Audible Analog Resolution to Your Digital Multimeter, Sam Green WØPCE, CQ Dec, pp. 18-22

ⁱⁱ Fun with Voltage-to-Frequency Converters, Sam Green WØPCE, QEX March/April 2013 pp. 7-10

ⁱⁱⁱ Microwave Version of Wideband QRP SWR Meter, Sam Green WØPCE, QEX January/February 2018 pp. 16-18

3D Printed Feedhorns for 10 GHz

Paul Wade, W1GHZ ©2019

www.w1ghz.org

3D printed horns have been shown to work quite well¹, and are not hard to make. The plastic horn is printed on a 3D printer, and then coated with metallic paint. There is a bit of technique involved in getting good conductivity, but Glenn Robb, KS4VA, described it well².

One problem is getting a coax connector attached to a plastic horn or waveguide. I have made circular waveguide transitions³ using ¾" copper water pipe, and made feedhorns by fitting a metal corrugated horn onto the copper pipe. Why not 3D print the complicated part, the corrugated horn, and fit it onto a copper pipe transition?

I reviewed my previous feedhorn simulations and selected corrugated horns which provided good feed performance: one for offset dishes and one for prime-focus dishes. I modified the dimensions slightly, increasing the wall thicknesses for better 3D printing. A quick simulation suggests that the performance is not changed significantly. STL files for both horns are available.

The metallized corrugated horn should slip over the copper pipe. The offset horn should be adjusted so that the rim of the pipe forms a corrugation at which lines up at the same angle as the others. The prime horn position is adjusted for different f/D – the corrugations should be nearly flush with the rim of the pipe for shallower dishes, and further back for deep dishes.

Getting the corrugated horn to slip over the pipe may take a bit of fiddling, as many 3D printers do not print to exact dimensions. For instance, my cheap printer prints about 2% undersize, so I adjust the scaling in the slicer program (I use CURA).

Many hams and most makerspaces have 3D printers. If you don't, there are also services that will print a file for a reasonable cost, and files for thousands of printable objects may be found at thingiverse.com.

Notes:

1. Paul Wade, W1GHZ, "3D Printed Horns," (Microwavelengths), *QST*, January 2019.
2. Glenn Robb, KS4VA, "3D Printing Microwave Antennas", *VHF Super Conference 2019 Proceedings*.
3. Paul Wade, W1GHZ, "Understanding Circular Waveguide—Experimentally," *QEX*, January/February 2001, pp 38-48.

The gate voltage, and the bias current, is set by the potentiometer R5, which has only a small adjustment range. As the NPN transistor heats up, it draws more collector current which pulls the gate voltage lower (Note: The NPN transistor must be connected – without, the gate voltage is significantly higher). The op-amp, U2, is a unity-gain amplifier to drive the gate. The op-amp was chosen by Ampleon to provide a low output impedance and because it is stable with a capacitive load (like the LDMOS gate).

The other potentiometer, R3a, can be used to set the voltage regulator output voltage. Since this voltage isn't critical, a fixed R3 resistor is adequate – if you are fussy, use the next higher value of R3 and a 100K pot at R3a. The Ampleon circuit set this voltage at 8 volts. Mike had some relays suitable for 8 volts, but Paul only had 12 volt relays so adjusted the bias circuit values for 12 volt operation. Resistor values for both options are in the parts list. The relay may also be powered separately, from terminal VR, after cutting the trace next to C3 at the point marked **x**.

Temperature compensation is set by R10. The 10K value shown is for the nominal $-2\text{mV}/^\circ\text{C}$. Reducing the value of R10 will increase the temperature compensation, and vice-versa.

Ampleon suggests an additional resistor, 5 to 20 ohms, in series with the LDMOS gate to ensure low-frequency stability. Adding a 5 volt Zener diode to ground might help protect the sensitive LDMOS gate from voltage and static spikes.

Figure 2 shows a prototype PC board mounted on an amplifier. Construction might be more elegant if it weren't an add-on.

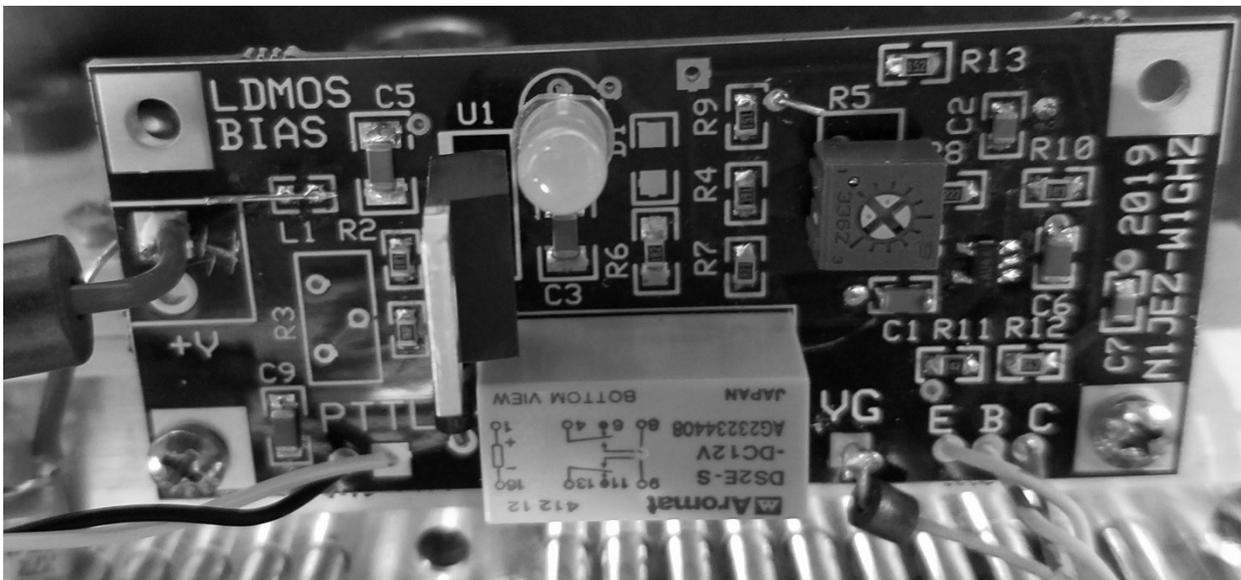


Figure 2 - LDMOS Bias Board Prototype mounted on amplifier

If anyone is interested, we can supply Gerber files for the PC board, which can be sent to OshPark (www.oshpark.com) to get 3 boards made at low cost. If there is sufficient interest, Paul might order a batch of PC boards.

LDMOS Bias Board

Parts List

<u>Designator</u>	<u>8 volts</u>	<u>12 volts</u>
C1	1uf	
C2	0.1uF	
C3	1uf	
C5	0.1uF 100V	
C6	1uf	
C7	0.1uF	
C9	0.1uF	
D1	LED	LED
L1	Ferrite Bead	Ferrite Bead
R2	180	180
R3	1K	1.5K
R3a	optional	optional
R4	390	390
R5	200 pot	200 pot
R6	1K	1.8K
R7	390	1K
R8	2.2K	2.2K
R9	75	75
R10	10K	10K
R11	1K	1K
R12	5.6K	5.6K
R13	1K	1.5K
Q1	2N3904	2N3904
U1	LM317AHV	LM317AHV
U2	LM7321MFX	LM7321MFX

Phase Noise Measurements of Some Synthesizers

Paul Wade W1GHZ ©2018, updated ©2019
w1ghz@arri.net

At the 44th Eastern VHF/UHF/Microwave Conference in April 2018, I got a chance to measure phase noise of some of the newer microwave synthesizers. Few hams have test equipment capable of making phase noise measurements of good oscillators, so we must rely on test equipment at various VHF and Microwave conferences. The excellent equipment at this conference was provided by Greg Bonaguide, WA1VUG, of Rohde & Schwarz.

More recently, at the VHF Super Conference in April 2019, I tested a new synthesizer, as well as one of the best previous ones for comparison. Since Greg provided different Rohde & Schwarz test equipment, the results are presented separately.

A frequency synthesizer is an attractive way of generating a signal at a desired frequency, particularly since it has become very difficult to find quality crystals. A modern synthesizer may operate at a high enough frequency to provide the Local Oscillator for a microwave system, and the frequency may be locked to an accurate reference to provide frequency accuracy and stability.

Phase Noise

However, the phase noise generated by almost all synthesizers is significantly worse than a good crystal oscillator. For very weak signals, my experiments¹ in 2009 suggest that the difference in Minimum Detectable Signal is about 2 dB, between a multiplied crystal oscillator LO and a synthesizer LO. On the other hand, 10 GHz MDS tests at the NEWS (North East Weak Signal group – www.newsvhf.com) picnic over several years suggest that knowing the frequency of a very weak signal can improve the MDS by up to 5 dB, when listening by ear. The addition of an SDR waterfall display eliminates the unknown frequency problem – all signals appear on the screen – so minimizing phase noise can help to hear very weak signals.

In 2012, I developed a locked VCXO² which can provide the source for a microwave LO with phase noise very nearly as good as a crystal oscillator and also be locked to a reference source to provide frequency accuracy and stability. This might be an ideal solution, but there are few choices for available VCXO frequency. As part of the development, phase noise comparisons were made with the synthesizers available at that time, shown in Figure 1. The bottom three curves clearly show how phase noise increase with frequency multiplication, whether done by a classic frequency multiplier or in a phase-locked loop. Since all the synthesizers are operating at 1152 MHz and are referenced from the same 10 MHz TCXO, any phase noise greater than the multiplied VCXO is additional noise generated by the synthesizer.

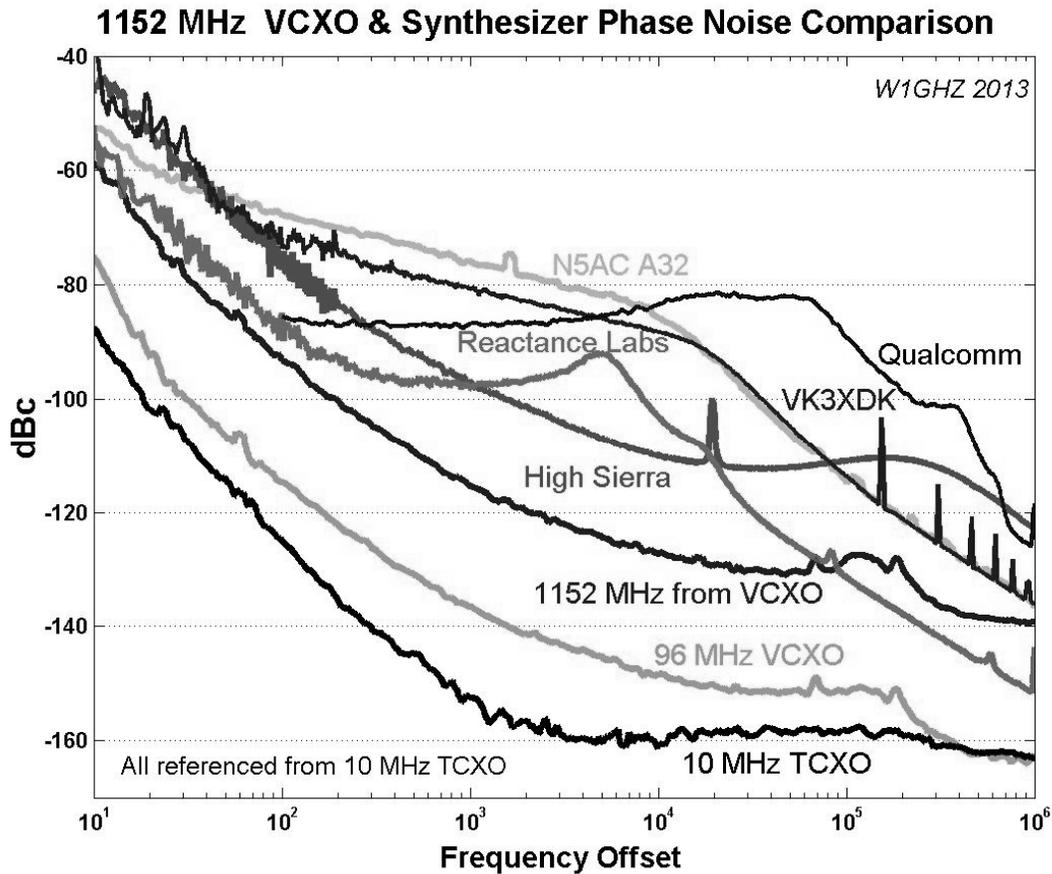


Figure 1 - Synthesizer Phase Noise measurements from 2012 and 2013

More recent synthesizers may have better phase noise characteristics than the N5AC A32 used for my 2009 experiments. Many of the synthesizers in Figure 1 are better, at least over part of the range of frequency offsets.

The measurements at the 44th Eastern VHF/UHF/Microwave Conference in April 2018 are shown in Figure 2. These were made with a Rohde & Schwarz FSW-43 analyzer. Where possible, synthesizers were referenced to a 10 MHz TCXO, the same unit as Figure 1. The VCXO system and several other synthesizers from Figure 1 are also included for comparison, including the N5AC A32. Some of the units were provided by conference attendees for a wider range of comparisons. Figure 3 includes a picture of each unit, and the Appendix provides more details.

Note that these are not definitive measurements, just what we able to accomplish during the lunch break at the conference. The measurements were made with 10X averaging, so they should be reasonably accurate. Better results might possibly be found with different programming of the synthesizer chips – for instance, for some New England beacons, W1EX found that an ADF4153 programmed for multiplication by four to 10368.320 or 10368.400 had fewer spurious outputs than at other nearby frequencies.

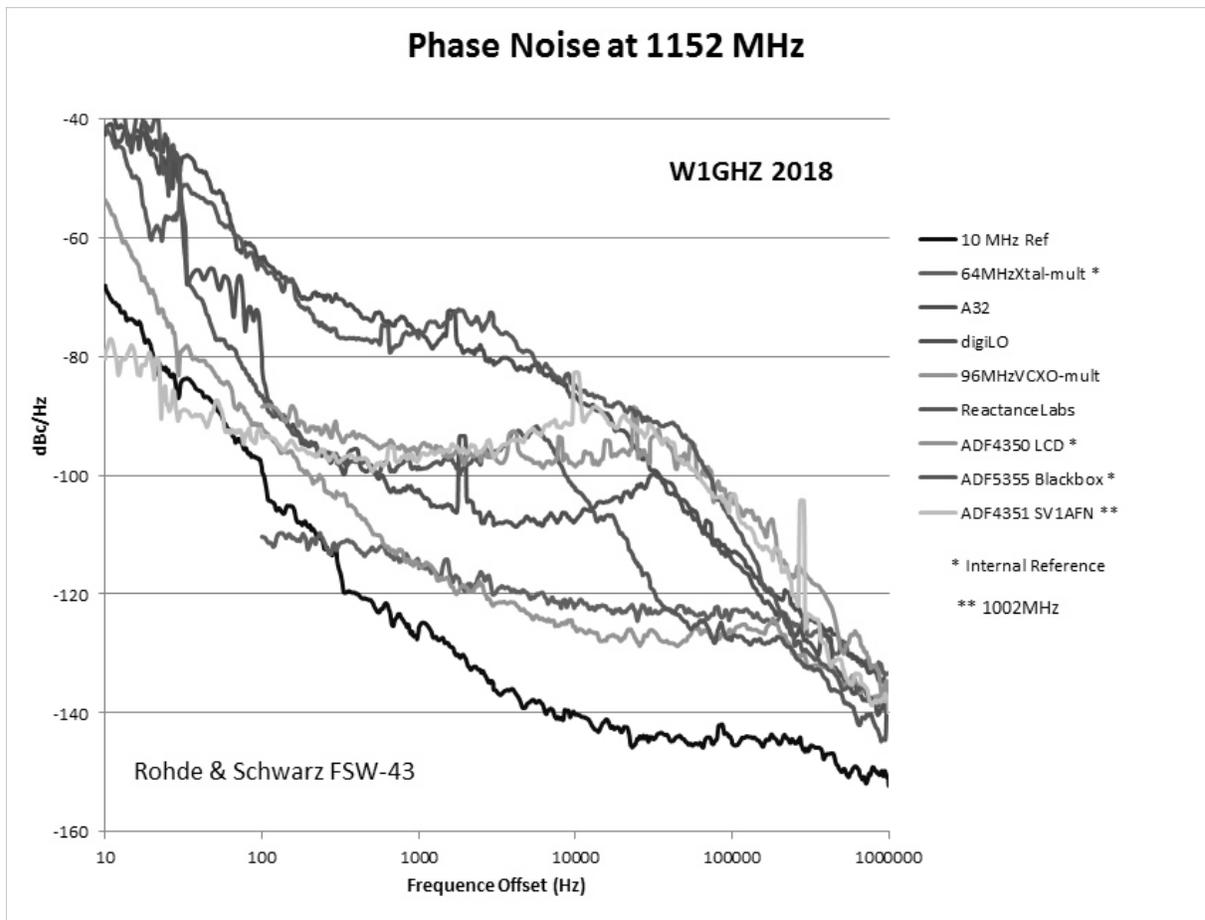


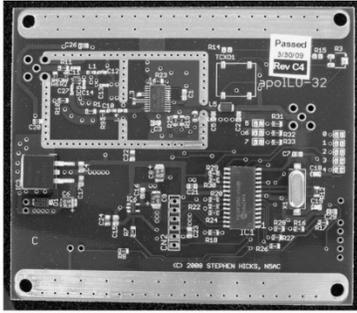
Figure 2 – 1152 MHz Phase Noise measurements at 2018 Eastern VHF Conference

Several of the newer synthesizers have pretty good performance. The digiLO from Q5 Signal (q5signal.com) has the best phase noise at 1152 MHz except for a spike at 2 KHz; it can be easily programmed with jumpers to popular ham frequencies from 23.5 MHz to 6 GHz.

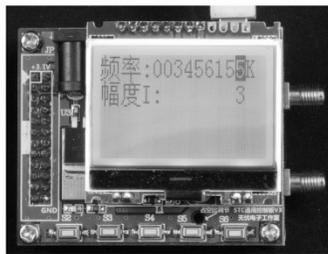
A surprisingly good one is the ADF4350 with the LCD display and programming buttons³, available from China on ebay, which goes up to 4 GHz. The buttons make it able to run standalone – with a USB battery, it makes a handy signal source.

And a good cost-effective one is the SV1AFN ADF4351 (www.sv1afn.com/adf4351m.html), which requires something like an Arduino for programming, to frequencies anywhere between 35 and 4400 MHz.

The ADF5355, available complete as shown or as a programmable board, operates up to 13.6 GHz. This one arrived just before the conference, so I didn't get a chance to check it out thoroughly.



A32 - N5AC



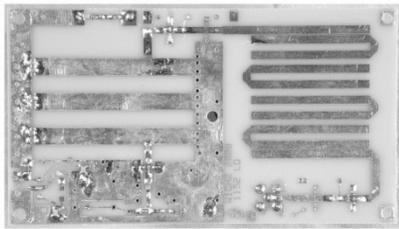
ADF4350



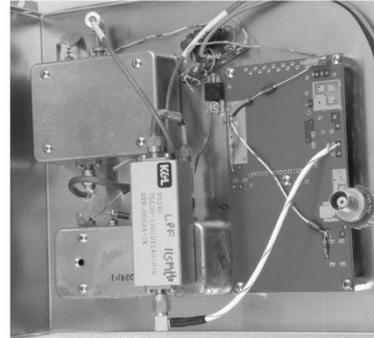
OLED digital ADF5355 54M-13.6...
ADF5355

digiLO
q5signal.com

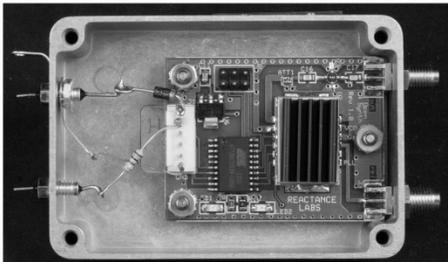
**1152 MHz
Sources**



64 MHz Oscillator & Multiplier



96 MHz VCXO & Multiplier



Reactance Labs

Floating-Ground by SV1AFN

**ADF4351 PLL
Synthesizer Module**

35MHz to 4400 MHz
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Make a VFO, Clock Gen etc. fast

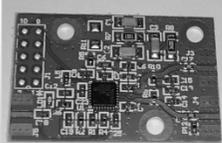


Figure 3 – 1152 MHz sources measured in Figure 2

X-Band Synthesizers

Conference attendees also brought several synthesizers that work directly at X-band, good for 10 GHz or higher bands. The phase noise of these units is comparable to what we would expect from the other synthesizers after frequency multiplication. The 120 Hz spike on the ZL-PLL curves are hum from a crappy power supply – a clean power supply makes a difference.

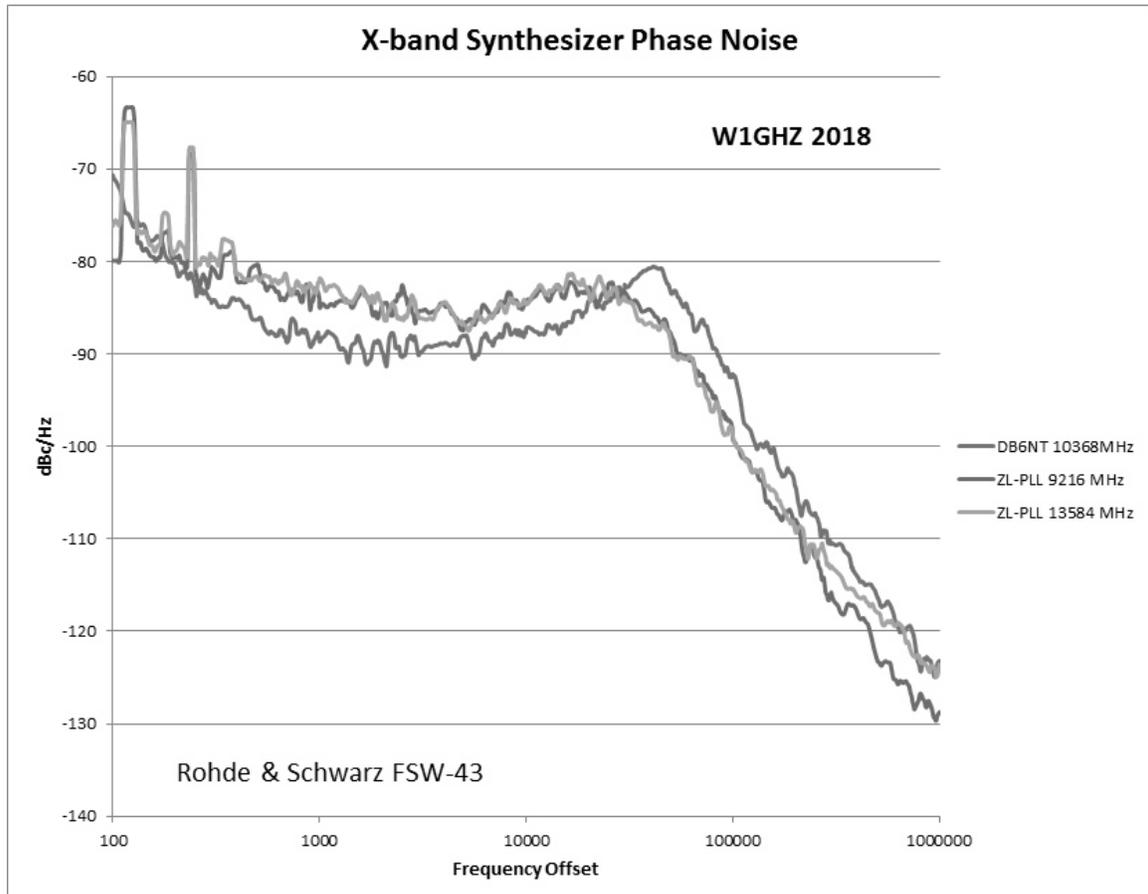


Figure 4 – X-band Phase Noise measurements at 2018 Eastern VHF Conference

DB6NT = MKU LO 8-13 (kuhne-electronic.de)

ZL-PLL = ZL-PLL 14G (zl2bkc.com)

2019 Measurements

Since the previous measurements, a new synthesizer has become available from VHFDesign.com, the LO-PLL-USB-MAX2871-SHF-PCB. This unit has attractive features: it is programmable from 1 to 6000 MHz, and includes a programmable beacon mode. The phase noise results at 1152 MHz are shown in Figure 5. Compared to the DigiLO, the best synthesizer measured so far, this unit is almost 20 dB worse, or comparable to some of the inexpensive synthesizers in Figures 2 and 3.

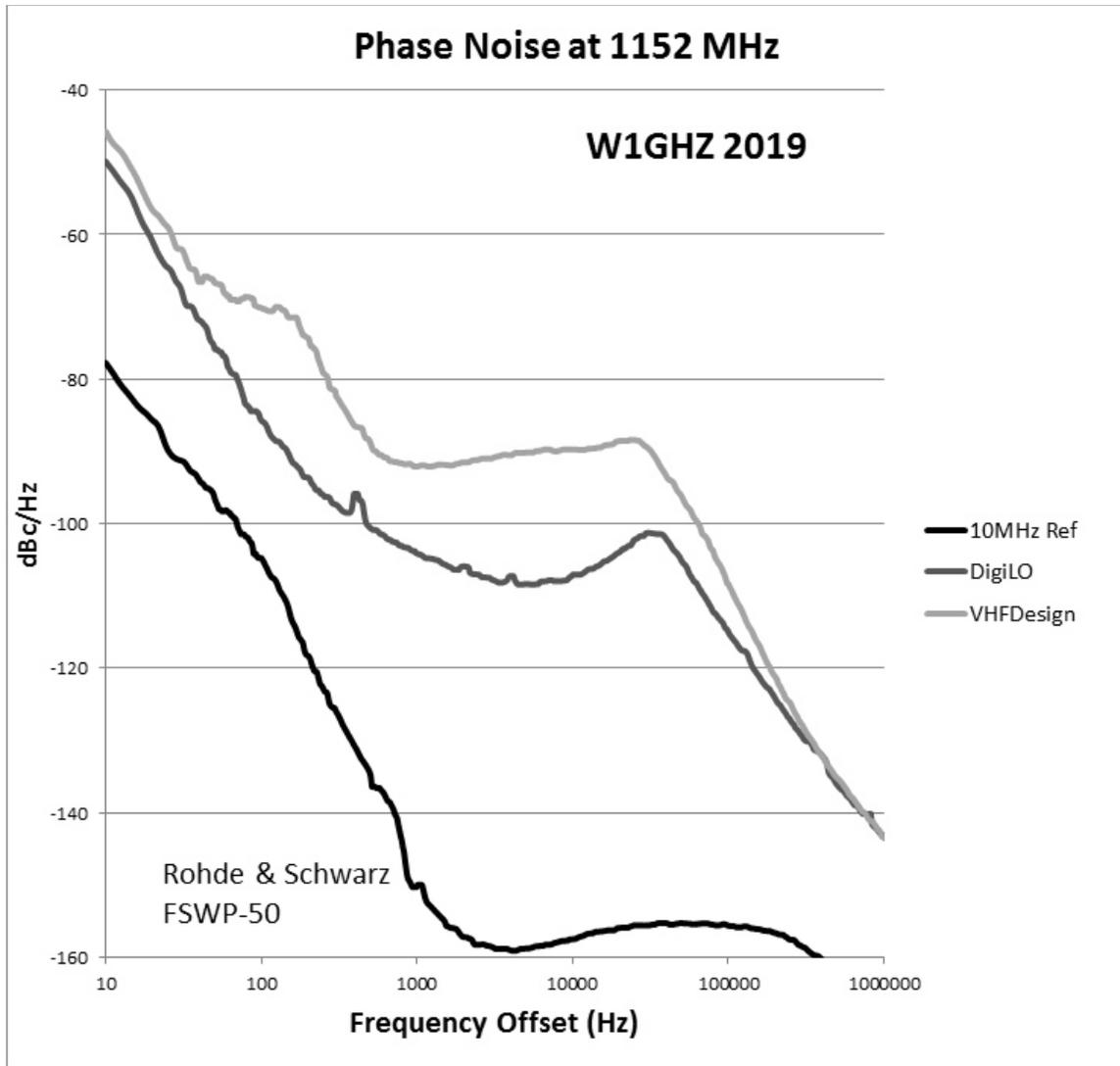


Figure 5 - 1152 MHz Phase Noise measurements at 2019 VHF Super Conference

Notes:

1. Paul Wade, W1GHZ, "Phase Noise and MDS," *Proceedings of Microwave Update 2009*, ARRL, 2009, pp. 193-196.
2. Paul Wade, W1GHZ, "A Flexible VCXO Locking Board," *Proceedings of Microwave Update 2012*, ARRL, 2012, pp. 101-113.
3. Paul Wade, W1GHZ, "Synthesized Signal Source From China," *44th Eastern VHF/UHF/Microwave Conference*, 2018.

76 GHz Transverter Upgrade Using the New DB6NT MKU 76 G2 Transverter

Al Ward W5LUA

Introduction

Last fall, several of us decided to make purchase of the new DB6NT transverter for 76 GHz. The new DB6NT design takes advantage of newer MMIC technology in this frequency range. Most notable is the use of image reject mixers for both receive and transmit. The new transverter also includes a PA that is specified at greater than 250 mW output. Noise figure is specified at about 8 dB which is still a considerable improvement over a simple passive diode mixer.

The 76 to 81 GHz Band

Europe has settled on 76032 MHz as the weak signal portion of the band while a good portion of the US has always been at 78192 MHz. In California, the chosen segment of the band has been closer to 80 GHz based on availability of components. Our EME activities of several years ago took place on 77184 MHz based on optimum performance of RW3BP's TWT. Recent regulations in the US have also limited our EIRP to 315W EIRP as we share the 76 to 81 GHz band with collision avoidance.

Integration

My first 76 GHz transverter used individual discrete components including homebrew tuned filters to cover the 77184 MHz segment that was used for initial EME tests with RW3BP[1]. The transverter used a 2m IF which necessitated very high Q filters with very narrow bandwidths. I was optimistic that the new DB6NT transverter would be able to cover a wider portion of the 76 GHz to 81 GHz of the band without the need to retune filters. Image reject mixers use unique phasing of the RF and LO signals to provide a nominal 20 dB of image rejection over some wider bandwidth than can be provided by filters thus allowing the use of direction conversion to 2m.

I started out by integrating the DB6NT transverter with a WA1MBA preamplifier and a WR-15 waveguide switch[2][3]. I also have a coupler in the transmit path to monitor power output. I use the DB6NT LO MKU LO 8-13 PLL with a power divider to feed the 76 GHz transverter and for possible higher bands.

The DB6NT LO has numerous preset frequencies including the 9486 MHz required for the 76 GHz transverter. The LO has the ability to store 2 additional user programmed frequencies. Since my original 76 GHz transverter was set up for 77184 MHz for EME tests with RW3BP, I decided to store 9630 MHz as F1. I also set up F2 as 9756 MHz to allow operation on 78192 MHz. With the proper "bridging" of programming pins on the DB6NT LO, I can switch between the 3 segments with a 3 position wafer switch.

A picture of my completed 76 GHz transverter is shown in Figure 1. Although the placement of the DB6NT transverter may seem unusual, my long term plan is to add a higher power amplifier which would push the transverter to the left. I would then add a corresponding piece of waveguide to the receive path to keep everything aligned. The transverter is assembled on a standard 10.5 inch by 19

inch (26.7 cm x 48.3 cm) rack panel with slots milled down the middle to allow the completed assembly to be installed on my 2.4m offset fed dish for moon noise and sun noise measurements.

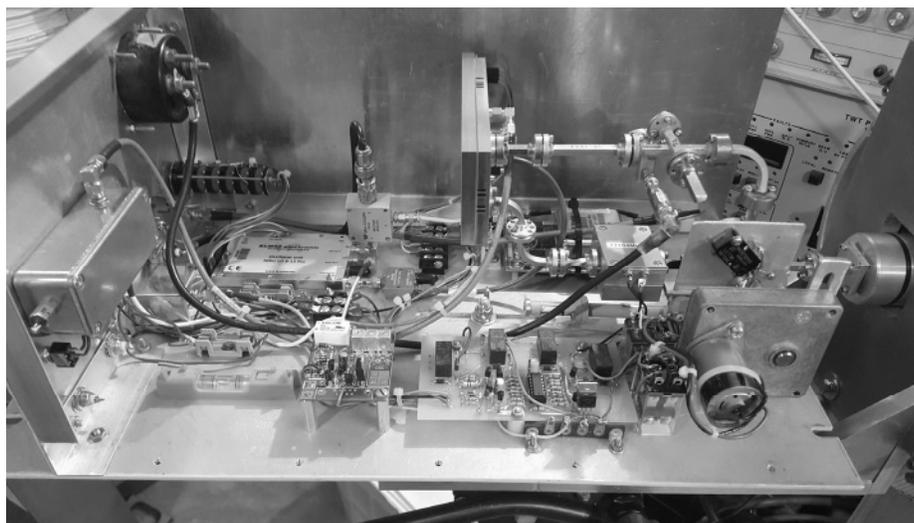


Figure 1 Completed 76GHz Transverter

Performance

I started out by testing the DB6NT transverter as a complete system on all 3 band segments. Keep in mind, the DB6NT transverter is only specified at 76032 MHz. I was interested in making 3 sets of measurements. Most importantly was measuring the output power followed by image rejection and then noise figure.

I used an HP432A power meter with a Millitech 45744H-1100 WR-15 head with a correction factor at 75 GHz. I also have a Quinstar QAF-E30000 30 dB WR-15 attenuator which I measured the loss as accurately as I could with my HP8757 scalar analyzer at the 3 frequencies of interest. At the waveguide relay antenna port, I was able to measure 275mW at 76032 MHz. Measuring the output power directly at the TX port of the DB6NT transverter provided the results shown in Table 1.

Freq (MHz)	Power output
76032	333 mW
77184	318 mW
78192	250 mW

Table 1 DB6NT transverter,
Power output vs Frequency

On the receive side, I measured noise figure with an HP8970B noise figure meter and a Noisecom model NC 5115 WR-15 noise source with a Hughes 45114H-1000 WR-15 isolator. The ENR has been established to be 15.3 dB at 77 GHz.

My DB6NT transverter is specified at a minimum of 28 dB conversion gain and noise figure less than 8.2 dB. Using the 2m IF port with transverter IF gain set at a maximum, I measure the performance shown in Table 2. As you can see the performance is very good.

Freq (MHz)	Conv Gain	Noise Figure
76032	28.1 dB	6.13 dB
77184	27.4 dB	5.75 dB
78192	26.9 dB	6.02 dB

Table 2 DB6NT transverter, 2m IF gain at max

Decreasing the DB6NT transverter IF gain to its minimum results in a transverter conversion gain near unity and the transverter noise figure will rise several dB. I then added in the WA1MBA LNA which has about 30 dB gain and then set the transverter IF gain at minimum with the results shown in Table 3.

Freq (MHz)	Conv Gain	Noise Figure
76032	31.7 dB	4.67 dB
77184	28.2 dB	4.09 dB
78192	26.7 dB	4.22 dB

Table 3 DB6NT transverter, w/ WA1MBA LNA
2m IF gain at minimum

Adding in the WR-15 waveguide relay completed the system. Table 4 shows the results.

Freq (MHz)	Conv Gain	Noise Figure
76032	31.5 dB	4.99 dB
77184	27.5 dB	4.69 dB
78192	25.4 dB	5.06 dB

Table 4 DB6NT transverter, w/ WA1MBA LNA,
WR-15 waveguide relay 2m IF gain at minimum

One of the major selling factors of the new DB6NT transverter is its' image rejection on both receive and transmit thereby minimizing the need for additional filtering in the RF paths. DB6NT specifies the image rejection to be 20 dB typical at 76032 MHz. For testing purposes, I use an ST Microwave 6139-6288-00 multiplier that has a female SMA on the input and a male 3.5 mm connector on the output. The module runs off of +8V. On 47 GHz it works nicely as a weak signal source when driven by my HP8340A in the 23.5 GHz frequency range. It is also capable of generating a nice 76 GHz weak signal when driven as a X3 multiplier from the HP8340A. In fact it generates a nice bench signal without the need for the +8V source. I simply put a 3.5mm to WR-28 adapter on the multiplier output and set it a few inches away from the waveguide switch opening. Crude but it works. I used my Elecraft KX-3 and P-3 Panadapter to make the measurements. I then generate the "real" frequency for a reference and then generate the "image" frequency. The real and image frequencies are separated by twice the IF or 288 MHz.

The results are very good as shown in table 5. The receive image rejection is greater than 20 dB over all 3 frequency bands. I have not looked at the image rejection on transmit but expect it to be very similar.

Freq (MHz)	Image Rejection
76032	26 dB
77184	30 dB
78192	22 dB

Table 5. DB6NT transverter Receive
Image Rejection

Other points of interest

The Monitor Output at Pin 9 on the transverter provides a convenient point to confirm one is putting out power. However it does not appear that the detected output voltage is connected to the forward port of a directional coupler as confirmed by the fact that the voltage changes with different loading on the output port. In my completed system, I measure about 1.9v at full power on 76032 MHz. The diode is most likely just measuring RF voltage on the output line.

The DB6NT LO provides a 116 MHz out at the “REF OUT” port and is intended to be used with the 2m to 10m mixer which is also included in the 76 GHz transverter. In the interest of minimizing the chance of “injury” to the 76 GHz transverter, I decided to use the “IF OUT” port at 2m and the low power option at the “IF/IN” port and drive the transverter with the W1GHZ Miniverter- F which Paul designed for the Flex-1500. I just use the 116 MHz LO to drive the Miniverter. Since the Elecraft KX-3 does not have a low power transmit port, I added a Downeast Microwave TC board at 28 MHz to interface the KX-3 to the Miniverter. If for some reason I run too much 10m IF power in, I will only harm the IF board and not the 76 GHz transverter. I call it cheap insurance. I also ran the PTT line out the IF port to the KX-3 using Bias Tees to minimize forgetting to hook up the PTT line.

The Test

After having spent years building equipment for 77 GHz which finally led up to my copying Sergei RW3BP off the moon, I had yet to make a 2 way QSO on 77 GHz. In March 2019, Tony K8ZR came down to Dallas and we both took off towards Phoenix, AZ where Barry VE4MA was spending his winter. On March 8, I had my first QSOs on 78192 MHz with VE4MA and K8ZR. We then moved to DM43ee and worked W7QQ and N0IO who were in DM33wo at a distance of 66km. The final test took place on March 9th, when VE4MA, K8ZR, and myself moved to DM42ok56ig atop Mt. Lemmon at 9000 ft. near Tucson while N0IO and W7QQ were set up in DM33rn26sp. We had 59 signals on SSB at a distance of 207 km. Success!

Conclusion

The DB6NT transverter has more than met my expectations. Its power output, image rejection and noise figure are superb.

Thanks to VE4MA and K8ZR for reviewing.

73

Al Ward W5LUA August 28, 2019

References

1. Al Ward W5LUA, “Analyzing EME Path Loss at 77 GHz”, Proceedings of Microwave Update 2014, ARRL, pp. 104-112.
2. Tom Williams WA1MBA, “78 GHz LNA Wrap-Up”, Proceedings of Microwave Update 2013, ARRL, pp 85-93.
3. Tom Williams WA1MBA, “78 GHz LNA Project Status Update”, Proceedings of Microwave Update 2012, ARRL, pp. 69-73.

Motorizing WR-10 & WR-15 Waveguide Switches

Al Ward W5LUA

Introduction

I acquired WR-10 and WR-15 manually operated waveguide relays and I wanted to automate their operation. The primary reason was to insure that my transverter was properly sequenced on transmit so as not to damage the LNA or the PA. I also wanted a quick way to automatically switch between cold sky and a termination. This sort of measurement is useful when making sky temperature measurements with respect to 290K.

Millimeter Wave Relay Characteristics

These millimeter wave type waveguide relays are a slightly different design than the customary 4 port designs used on the lower microwave bands. The lower frequency 4 port designs rotate 90 degrees to change states. These millimeter wave relays obtain a different position every 45 degrees of knob rotation. Besides the normal 90 degree bend in the relay, there is a straight through port that was once used for a “loop back test” that allowed the transmitter to be injected directly into the receiver for a “built in test function”. The configurations we will most likely use are shown in Figures 1 and 2. Figure 1 shows a straight through path between the receiver and the antenna. Figure 2 shows a 90 degree path from the transmitter to the antenna and a 90 degree path from the receiver to a port that is terminated in a load.

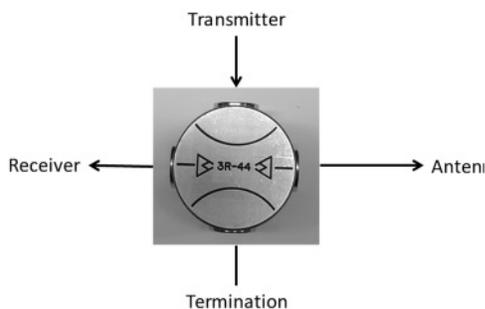


Figure 1 Waveguide Relay in Receive Position

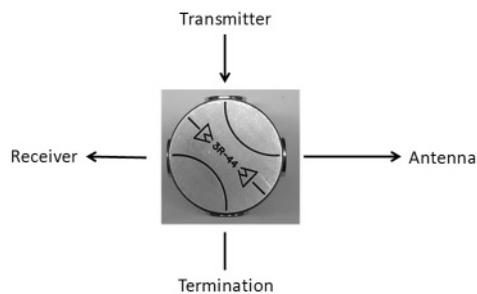


Figure 2 Waveguide Relay in Transmit Position

Motorizing the Manual Waveguide Switch.

Based on the desired configurations shown above, the waveguide relay shaft needs only to be moved 45 degrees between receive and transmit states. I acquired a small reversible 12vdc motor that travels at about 15 rpm. Limit switches will be used to stop and reverse the motion of the motor. After removing the knob from the waveguide switch, I inserted a long #4-40 screw through the shaft. Depending on your style of waveguide relay, you may or may not have to drill a hole to thread the #4-40 screw into. The #4-40 screw acts as a lever to actuate the micro-switches.

The micro-switches are mounted on an aluminum plate which is attached to the waveguide relay. The switches are oriented such that when the switches are fully compressed with the lever, the waveguide relay is either in the receive mode shown in Figure 3 or the transmit mode as shown in Figure 4.

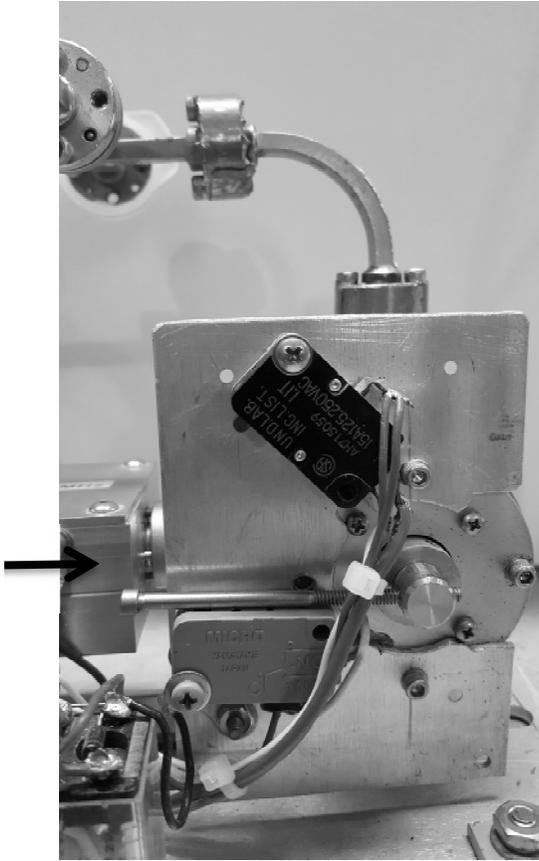


Figure 3 Waveguide Relay in the Receive position

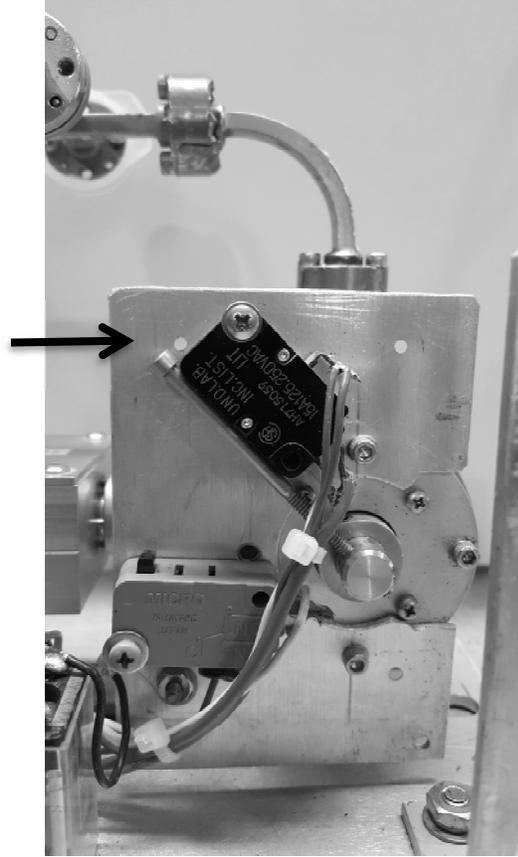


Figure 4 Waveguide Relay in the Transmit position

The switches and motor are wired as shown in Figure 5. Switch S1 is the upper switch and switch S2 is the lower switch as shown in Figures 3 and 4. The normal “resting” state is in the receive mode where S2 which is a normally closed switch is held open to leave the motor un-energized. On transmit +12V from the sequencer board is applied to the “+12V in to XMIT” port. This energizes RY1 and the motor causing the motor to rotate towards the transmit position. Once the motor has moved the waveguide relay to the transmit position, the lever compresses S1 causing the motor shut to shut off, even though RY1 is still energized. Only after the waveguide relay is fully in the transmit position, is the voltage “12V out on XMIT” available. I use this voltage to control another 12v relay that has a NO contact that pulls the PTT line of the transverter to ground. The transverter is guaranteed not to go to transmit unless the waveguide relay is fully in the transmit state.

When the sequencer board removes voltage from the “+12V in to XMIT” port, then relay RY1 is de-energized and the motor moves the waveguide relay toward the receive position and when the lever hits S2 again the motor shuts off and the waveguide relay is back in the receive position.

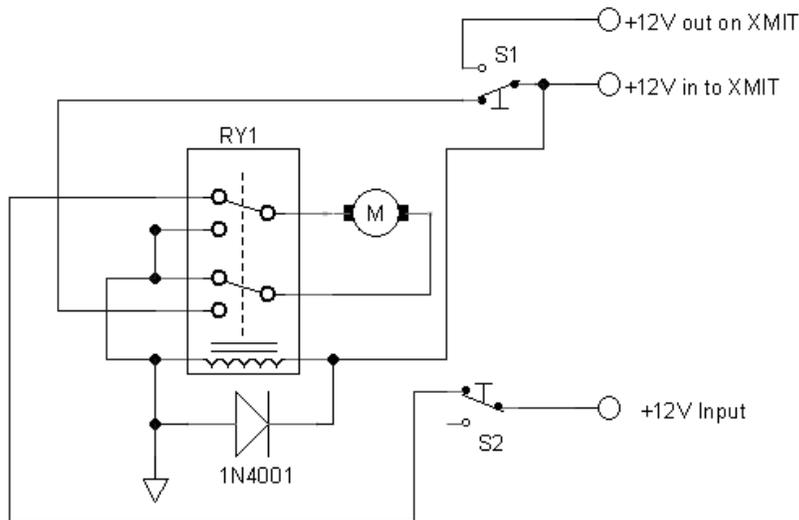


Figure 5 Schematic diagram of relay switching circuit

Figure 6 shows a top view of the completed motorized waveguide switch including the motor and shaft coupling.

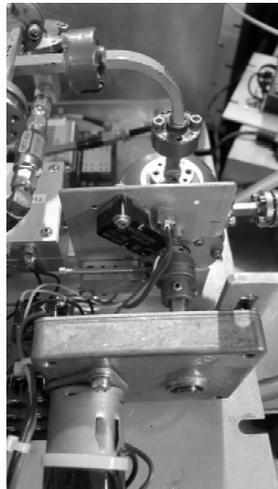


Figure 6 Motorized waveguide relay

Summary

A simple approach to motorizing a millimeter wave switch has been described. The circuit also provides a safety interlock to make sure the transverter or PA will not go to transmit if the waveguide relay is not fully engaged in the transmit position.

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Al Ward, W5LUA

August 28, 2019

RelCom WR-28 Waveguide Switch at 47 GHz

Al Ward, W5LUA

As amateur radio operators, we tend to make the best use out of what we have to work with. This is especially true when we use waveguide or waveguide components that were primarily designed for a band outside our frequency of interest. On 10 GHz, where the primary waveguide size is WR-90, we quite often use the next sizes up like WR-75 and WR-62. On 24 GHz, WR-42 is the primary waveguide of choice but we do use WR-28 which is designed for 26 to 40 GHz. The low frequency cutoff of WR-28 is 21.081 GHz allowing the waveguide to pass 24 GHz with little or no attenuation. Moving up to 47 GHz, we would normally be looking for WR-22 or WR-19, but this waveguide gets rather pricey. So why not try WR-28? Especially if you are looking for a waveguide switch for 47 GHz.

When Greg AA5C was assembling his 47 GHz transverter, I suggested that he try to use the RelCom WR-28 waveguide switch as a TR switch. I was unaware of the magnitude of the problem that Barry previously described at 47 GHz with the RelCom waveguide switch. Measurements by Greg on his 47 GHz transverter with the RelCom relay installed showed a 4 dB plus increase in receive noise figure and a 4 dB loss in power out put on transmit. Setting up bench tests, it became evident that this was truly the case. These observations were predicted by Barry in his swept measurements. It was also apparent that intentionally mis aligning the flanges could lower the loss.

Our attention then turned to the choke that was machined into the RelCom switch. A cross sectional view of a typical choke to flat waveguide junction is shown in Figure 1.

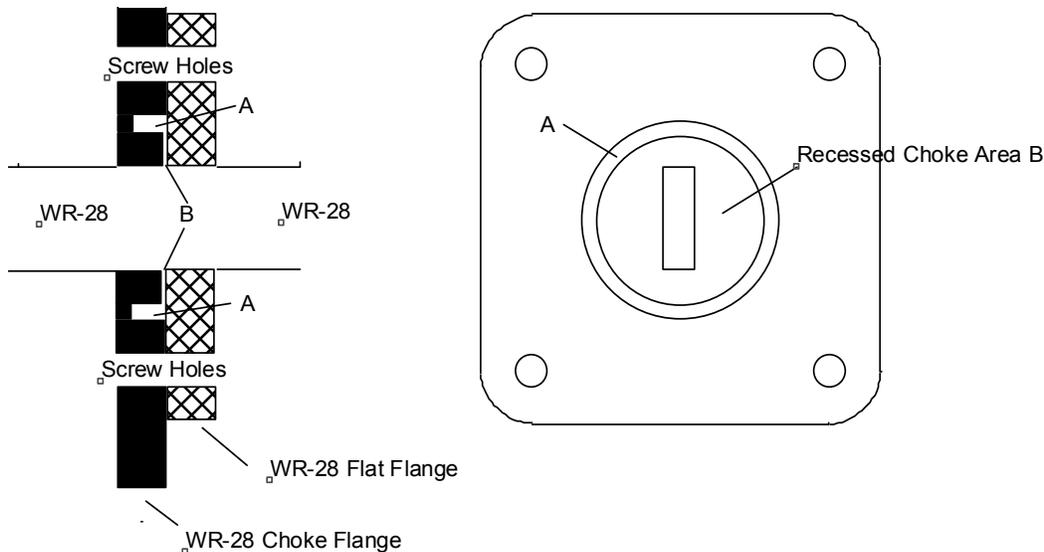


Figure 1. Cross sectional view

The choke consists of 2 resonant structures A and B. The groove at A is roughly a quarter-wave long at the original design frequency. B shows a recessed area that is inside the area defined by the groove A. Both of these structures are designed to provide a low impedance at the junction of the choke flange and the flat flange. This helps minimize loss at the junction of the 2 waveguides. Since WR-28 is normally optimized for a frequency in the 26 to 40 GHz frequency range, it is not surprising that the WR-28 choke might have some unusual response at 47 GHz.

At this point, I decided to compare WR-28 choke flanges. I went into my waveguide box and found what I believe to be a typical WR-28 flange with choke and I compared dimensions between this choke and the choke on the RelCom switch. The comparison is shown in Figure 2.

Parameter	Choke Depth	Choke Width	Recessed Area
Std Flange	.100"	.022"	.006"
RelCom Flange	.100"	.025"	.0025"

Figure2. WR-28 Choke Flange Dimensions

It was apparent that both of these flanges probably have different resonant frequencies or traps that may affect 47 GHz. I had to find a way to change the resonant frequency of the choke. I decided to fill the trench with something conductive. I used the leads from a ¼ watt carbon resistor whose wire diameter was .025". I was able to insert 2 layers of resistor lead in the trench on 2 ports of the switch. When retested, I had reduced the loss of the waveguide switch from nearly 5 dB, down to about 2.5 dB at 47088.1 MHz. Clearly the choke is having a significant effect on 47 GHz. Now comes the question of what can be done about the recessed area. It is only recessed by .0025" but encompasses an area of .324" and is still forming a sort of tuned circuit at the junction of the 2 flanges. I don't have a good answer other than trying some tuning on the WR-28 waveguide leading into the switch. If I was having an issue with a normal piece of WR-28 waveguide with a choke flange, I would file it flat as Barry as suggested. However, I would not chance filing on the RelCom switch. Put some screws in the waveguide and tune it up!

Bottom line...be aware of the WR-28 choke flange at 47 GHz!

Al Ward W5LUA August 29, 2019

Flyswatter Experiments at 47 and 77 GHz

Al Ward W5LUA

Introduction

I have always been intrigued by the flyswatter or periscope antenna ever since a local friend of mine, Craig Young, KA5BOU installed one on his tower for 10 GHz operation. Recently, he upgraded his to a different sized reflector and I became the owner of his original “stop sign” reflector. My goal was to install it on my Rohn 45 rotating tower primarily for 47 GHz and higher operation. It has been used successfully on 47 GHz for the last couple of years and I was anxious to try it on 77 GHz as well.

Concept

The beauty of using a rotating tower for the periscope antenna is that both the periscope reflector and the equipment rack located near the base of the tower are rotating in sync. This assures that the polarity of the ground mounted antenna is not rotated with respect to the reflector. Otherwise the polarity of the ground mounted antenna would have to be rotated to maintain the desired polarity. The periscope reflector that Craig built uses a legally obtained stop sign hinged for elevation or tilt control. A small actuator is used for adjustment on elevation. Craig used a small digital elevation indicator that he found at an RV shop. It only reads to the nearest degree which was probably ok for 10 GHz. A revised elevation scheme is on my list of things to do.

A big advantage of the flyswatter arrangement is that all delicate and pricey microwave equipment is ground mounted. Although, I normally do not leave the equipment at the base of the tower when it is not in use. This new taller tower seems to be a lightning magnet anyway.

Things to consider

Paul Wade's On-Line antenna book[1] has a downloadable excel spread sheet that allows one to easily optimize the major parameters in a periscope antenna system. On my tower I had already determined that the maximum separation between the reflector and the ground antenna would be 55ft or 16.8m. I chose this height based on the fact it was just below my second guy ring and I wanted all the larger microwave antennas to be between the second and third guy ring. I was also stuck with using the same reflector aperture of about 30 inches or .76m on all frequencies. My only real variable to optimize for a given band was the aperture of the ground mounted antenna. A summary of my predicted system performance is shown in Figure 1.

	10.3 GHz	24 GHz	47 GHz	76 GHz	Actual system
Frequency	10.3 GHz	24 GHz	47 GHz	76 GHz	76 GHz
Optimum Dish Diameter	.85m (33.6")	.5m (19.7")	.23m (9")	.15m (6")	.34m (13.5")
Flyswatter Aperture	.76m (30")				
Reflector Spacing	16.8m (55ft)				
Suggested Flyswatter Aperture	1m (39.4")	0.6m (23.6")	0.5m (19.7")	0.4m (15.7")	0.4m (15.7")
Final Results					
Dish Gain	36.6 dBi	39.4 dBi	38.5 dBi	38.9 dBi	46.1 dBi
System Gain	34.7 dBi	39.8 dBi	34.7 dBi	34.3 dBi	34.2 dBi
Effective Gain of Periscope over Dish	-2.0 dB	.4 dB	-3.8 dB	-4.6 dB	-11.8 dB
Figure 1 Summary of data from Paul Wade's on-line antenna book					

For my fixed reflector height of 16.8m (55 ft), the .76m (30") flyswatter aperture is probably optimum for 24 GHz with only slightly reduced performance at 10 GHz. At some point, I may actually try the flyswatter at 24 GHz and see how it compares to my Macom 2 ft dish at about 60 ft. However, my main goal was to experiment at 47 GHz and 76 GHz. It is interesting to note that with the optimum choice of spacing and aperture size, one can actually get some system gain from the flyswatter.

Results at 47 GHz.

To date, I have only run tests over a nearly unobstructed path of 23km (14 miles) from my home to an open spot on the other side of Lake Lavon in Farmersville in EM13td. This is based on an antenna height of 55 ft at my station. I have worked AA5C, AA5AM, and K8ZR during the 10 GHz and up contest through 47 GHz.

I attempted to document received signal strength from AA5AM during one of our earlier tests. Scott is running a DB6NT transverter at about -7dBm into a homebrew horn with a theoretical gain of 24 dBi assuming 50% efficiency. His EIRP is 17dBmi. The path loss is calculated from $20 \log(4\pi d/\lambda)$ and turns out to be 153dB. This does not account for any additional atmospheric loss from moisture, etc. The received signal at W5LUA should be -136 dBm. With my assumed system gain of a little more than 34 dB, the signal level at my receiver should be -102 dBm. Based on a 30 Hz effective bandwidth of the panadapter on my Flex-1500 I believe my noise floor to be $-174 \text{ dBm} + 5 \text{ dB NF} + 10 \log(30\text{Hzbw}) = -154 \text{ dBm}$. Scott's signal should have been 52 dB over the noise under perfect conditions. I measured his signal at 43 dB S/N ratio. The 9 dB difference is either in dish efficiency and or propagation anomalies. This test was run on Nov 29, 2017 when the temperature was 55F and the dew point was 45F.

Moving on to 77 GHz

On September 4, 2019, AA5AM and I attempted 77 GHz along with 47 GHz at the same EM13td site in Farmersville. The rig that Scott was using is the same rig that I used to copy RW3BP off the moon on 77184 MHz. Noise figure is sub 5 dB. The rig has since been modified on transmit as the output power is low, about -1 dBm at 77184.1 GHz. The antenna we started with on this rig was a small 1.1" X 1.4" commercial horn with a calculated gain of 26 dBi at 50% efficiency. Scott's EIRP would be about 25 dBmi. Path loss at 77 GHz is 157 dB plus 18 dB of additional path loss due to a Dewpoint of

68F. The additional path loss due to dewpoint was extrapolated from a curve created by Brian Justin WA1ZMS. The power level received at my location should be about -150 dBm. With my projected system antenna gain of 34 dB, the power level incident to my receiver should be -150 dBm plus 34 dB = -116 dBm. The sensitivity of my system with KX-3 with PX-3 panadapter should be -174 dBm + 5 dB NF + 10 log (11HzBW) = -158 dBm. I should receive the signal at 42 dB S/N on the PX-3 panadapter in a 5 kHz wide span. The 11 Hz panadapter bandwidth is based on information in the Elecraft PX-3 manual.

We had success on our first attempt. After some peaking on both ends, Scott's signal was about 22 dB over the noise at my home location. See Figure 2.

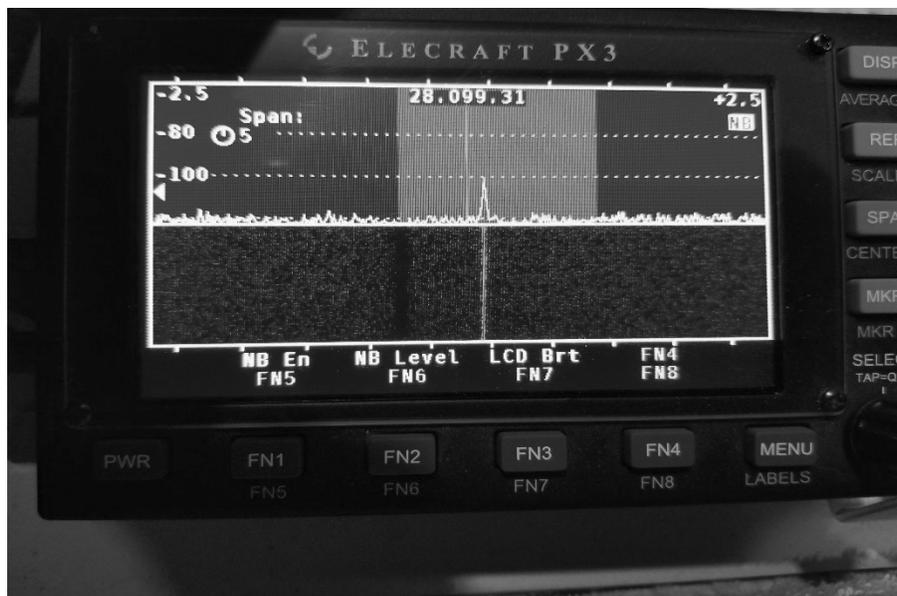


Figure 2. W5LUA reception of AA5AM who was using a small horn

Although the actual test result was 20 dB worse than predicted, I was still happy with the Flyswatter antenna system. I was not able to optimize my elevation due to an indicator malfunction. The Flyswatter was initially set at 45 degrees with respect to earth at this azimuth angle so it is close to being optimum.

The next transmission from Scott on 77 GHz was with a 6 inch diameter lens antenna loaned to the cause by Kent WA5VJB. The lens antenna was manufactured by Flann Microwave with a posted gain of 37.5 dBi but no mention of frequency on the label. It was a WR-15 horn so it may have been designed for 60 GHz. Figure 3 shows the enhanced signal as received by W5LUA.

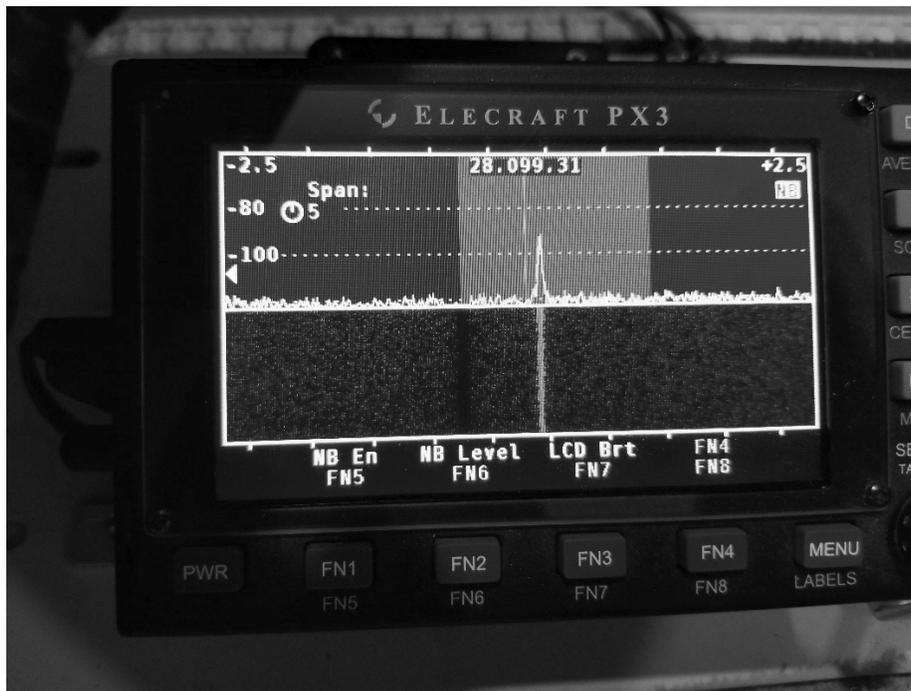


Figure 3. W5LUA reception of AA5AM who switched to a 6 inch Lens Antenna

Theoretically the 6 inch lens antenna should have a gain of about 12 dB over the small horn. Measurements show closer to 10 dB over an actual path. There was considerable fading on signals. Since we were shooting across our small Army Corp of Engineers Lake, there may have been some effects from the lake. As mentioned before the Dew Point was a comfortable 68F.

The next test for comparison was to try 47 GHz over the same path. Scott fired up the DB6NT transverter with a mighty -7dBm into a 24 dBmi horn and I saw similar S/N ratio at 47 GHz. I was able to mount both my 47 GHz transverter and the 77 GHz transverter on the platform at the same time. This made band switching easy.

Equipment Platform

I constructed my equipment platform out of two 6 ft. pieces of .25" thickness 2" aluminum angle anchored with u bolts to the Rohn 45 tower. The spacing between the two pieces of aluminum angle is 10.5". This works out to be convenient for me as both my 47 GHz and 77 GHz transverters are built on standard 10.5" high standard 19" rack panels. I add the equivalent of wings on each of the transverters to allow them to slide along the aluminum angle pieces. The center of the flyswatter reflector is 34 inches away from the tower. When I was first running tests on 47 GHz, I found the actual position for optimum S/N to be fairly broad. Possibly because I am illuminating a smaller portion of the flyswatter reflector. Although maybe not optimum, I decided to take advantage of this, and mount both rigs on the platform to allow for easy band switching. The final configuration is shown in Figure 4. The 47 GHz transverter is the one closest to the tower. The 77 GHz transverter has a Trimble GPS 10 MHz reference oscillator built in. Notice the GPS antenna to the left of the 77 GHz transverter. At the moment I am turning the tower with an HDR-300 rotor driven by a Green Heron Engineering rotor control box. Eventually, I will turn the tower with an M2 Orion rotor which will have better pointing accuracy.

The real challenge is to perfectly aimed each antenna at the flyswatter reflector. I use 2 guy wires and turnbuckles for support and aiming. I also use shims to perfect the alignment. A Sears torpedo level with digital readout helps as well.



Figure 4. Equipment platform with both 77 GHz and 47 GHz transverters mounted

A view of the reflector is shown in Figure 5. The reflector is about 30 inches at its widest and is spaced 34 inches out from the tower. The azimuth is fixed to hopefully keep it in perfect alignment with the equipment platform. It is variable in tilt only with a small actuator. When Craig was using this reflector he used a small rotator to rotate the flyswatter reflector as his tower was not a rotating tower.



Figure 5. Flyswatter reflector with elevation/tilt control.

Conclusion

The flyswatter reflector has more than met my expectations. As the results show, I have a few more dB to squeeze out of the system. I am hoping to work some greater DX and a few grids without going to a mountain top. As noted earlier in the analysis, the flyswatter reflector is too big for optimum performance at 47 and 77 GHz. I need to decide if it is worth going to a smaller reflector and forget using it also on 10 and 24 GHz. However, the number 1 item for me to work on is to fix my elevation indicator.

The upper bands are a real challenge and a lot of fun!

73 de Al Ward W5LUA September 5, 2019

Reference

1. W1GHZ Online Microwave Antenna Book. <http://www.w1ghz.org/antbook/contents.htm>

A low effort .1 – 2.5 GHz dBm power meter

Chuck MacCluer, W8MQW

w8mqw@arrl.net

08-14-2019

A simple dBm meter

In the event that you missed Andy Nehan's clever RadCom article on a simple UHF AD8307-based power meter ¹, I present here an even simpler AD8313-based meter for SHF. See Figure 1.

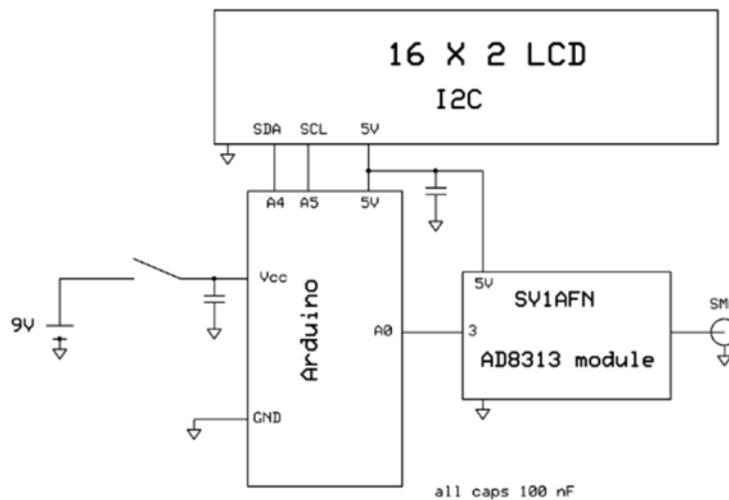


Figure 1. An Arduino reads the DC output voltage of the AD8313 logarithmic detector, then translates to dBm to be displayed on the serial LCD module.

I have also built this meter using lower cost AD8313 modules available on Ebay, but SV1AFN's high quality module is more stable. The dynamic range of this meter is from -60 dBm up to about 10 dBm. The modules report out a voltage of about 0.5 volts at -60 dBm, increasing linearly (with dBm) to 2.5 volts at 10 dBm. The Arduino does the A-to-D conversion, translates voltage to dBm, then displays the result through serial I2C communication on a I2C 16X2 LCD.

Construction

This is a very low-effort project when you use a transparent acrylic enclosure (from say *The Container Store*) since you merely glue the LCD panel to the inside front wall with four tiny dabs of Loctite super glue. Drilling the acrylic for the SMA and on-off switch is a matter of seconds. The 9-V battery and

¹ Andy Nehan, G4HUE, "RF meter for low power using an Arduino and LCD," RSGB RadCom, December 2018, pp. 75—77.

Arduino Nano are secured with hook-and-loop strips, and all connections are made with 4-inch breadboard jumper wires. See Figures 2 and 3.

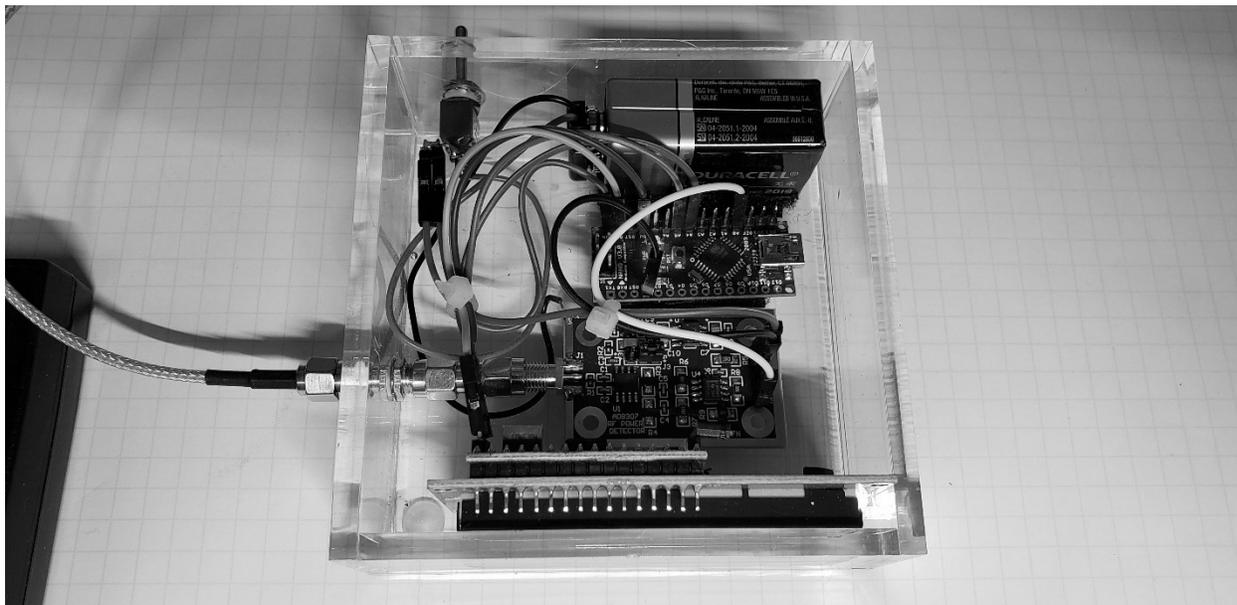


Figure 2. The LCD is spot glued to the inside front panel of the transparent acrylic enclosure while the Arduino and battery are secured with hook-and-loop strips. Wiring is quickly done by pushing on breadboard jumpers.

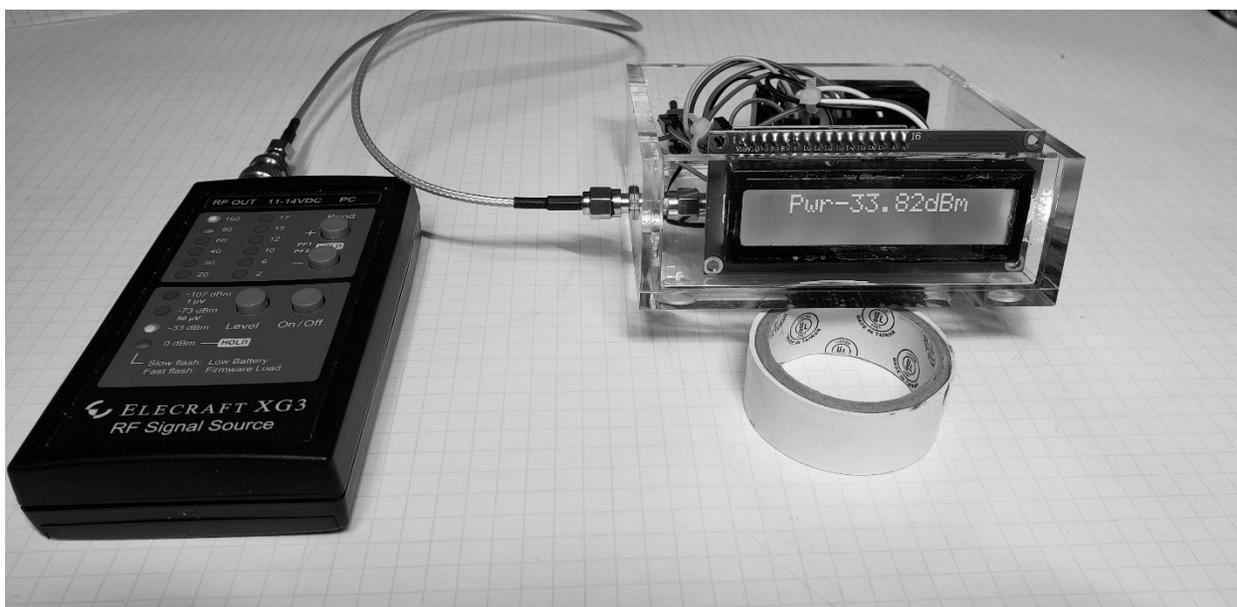


Figure 3. The power meter in operation.

Arduino sketch

I will be delighted to supply the code for this power meter in two flavors: The first is a one-pass A-to-D capture of the AD8313 output, the second employs averaging. The averaging seems necessary for the low-cost Ebay AD8313 modules, while the single pass appears sufficient for the SV1AFN module. Request by email at w8mqw@arrl.net.

You will have to calibrate the meter via the software for your particular module. You can experimentally tweak the values of the function *scaledb* found in the last two lines of the sketch until you are within acceptable accuracy. Or, more systematically, measure the output DC voltage of the AD8313 module at a high known power level (like 0 dBm), insert attenuation, then measure output voltage again at a low level (say -40 dBm), then apply the point slope formula for a line to obtain the exact slope and intercept for the translation from voltage to dBm.

Several lines of the sketch are devoted to aural readout, which you may ignore or delete. I needed the audio feedback once while adjusting an EME dish feed for minimal return loss and illumination. The audio quality of the Arduino's pulse-width modulation scheme is ghastly.

Quick Beacons

Chuck MacCluer W8MQW

w8mqw@arrl.net

08/14/2019

Although it is trendy to establish digital propagation beacons transmitting MSK144 or WSPR, the traditional CW beacons still play an important role --- especially at VHF, UHF, and SHF --- where such beacons are used to verify frequency or antenna-pointing accuracy and changes to preamplifier noise figure or overall system noise. Such CW beacons require a signal source, a beacon keyer, and an amplifier, as in Figure 1.

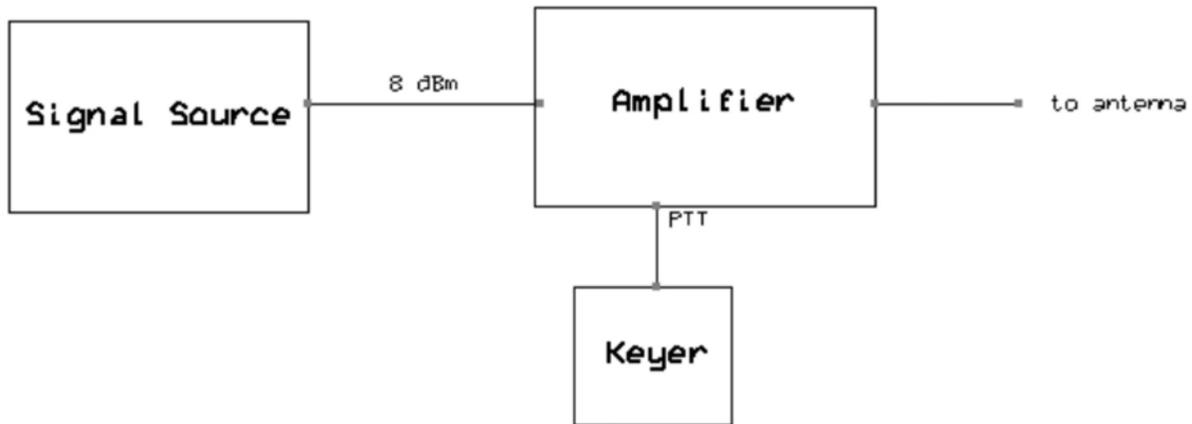


Figure 1. A typical CW beacon consisting of a signal source, a keyer, and an amplifier.

In the antediluvian past, a signal source at say 1296 MHz consisted of a 108 or 144 MHz crystal oscillator overdriving a MMIC with the 12th or 9th harmonic selected by a cascade of printed hairpin filters interspersed with MMIC amplifiers. But nowadays we have the Si570 from Silicon Labs and our signal source becomes simple indeed---see Figure 2.

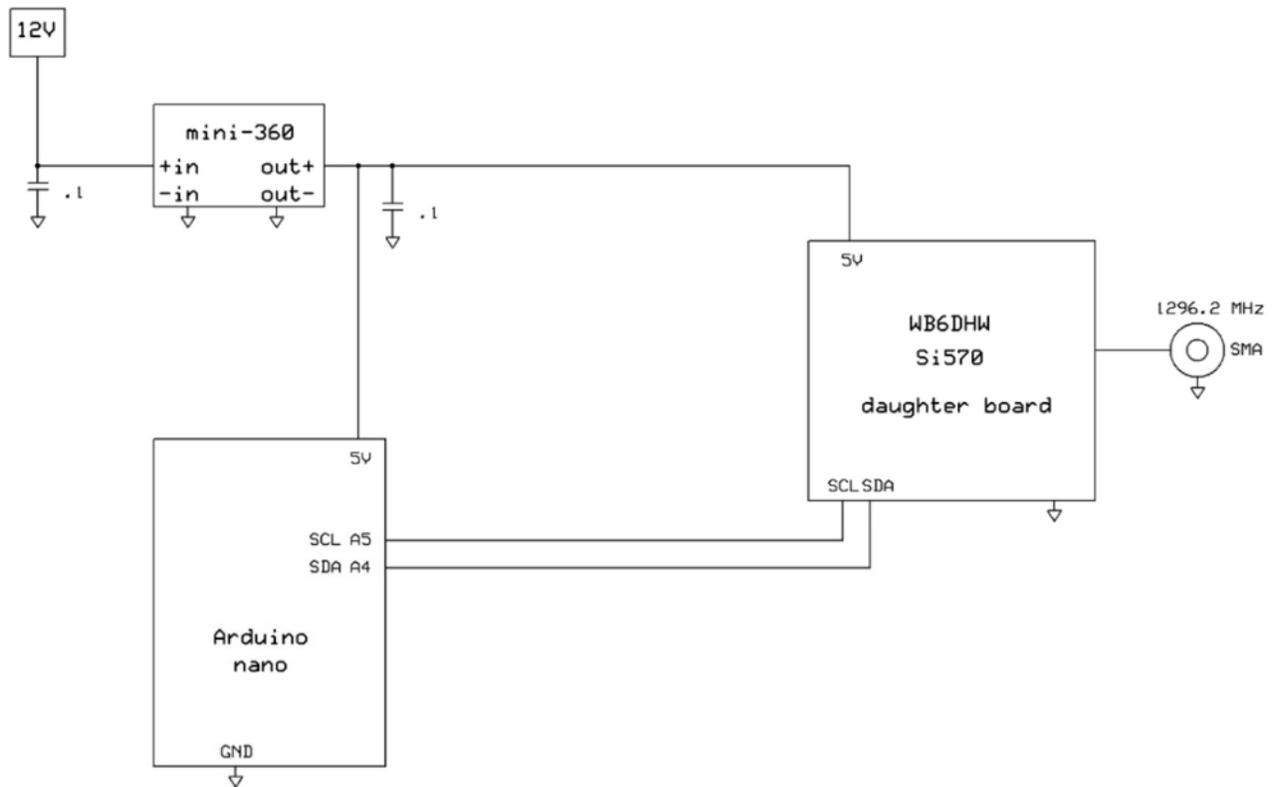


Figure 2. A simple beacon signal source using an Arduino (UNO, Nano, etc.) to write the frequency-determining registers of a Si570 programmable oscillator upon startup. The daughter board contains a Si570, a 3.3-V regulator, several pull-up resistors, and a level shifter to communicate I2C data from the 5-V Arduino to the 3.3-V Si570.

The Softrock Approach

For signal sources for beacons below 220 MHz there is an even simpler solution --- purchase a Softrock RX ensemble kit and use only the DDS portion as in Figure 3.

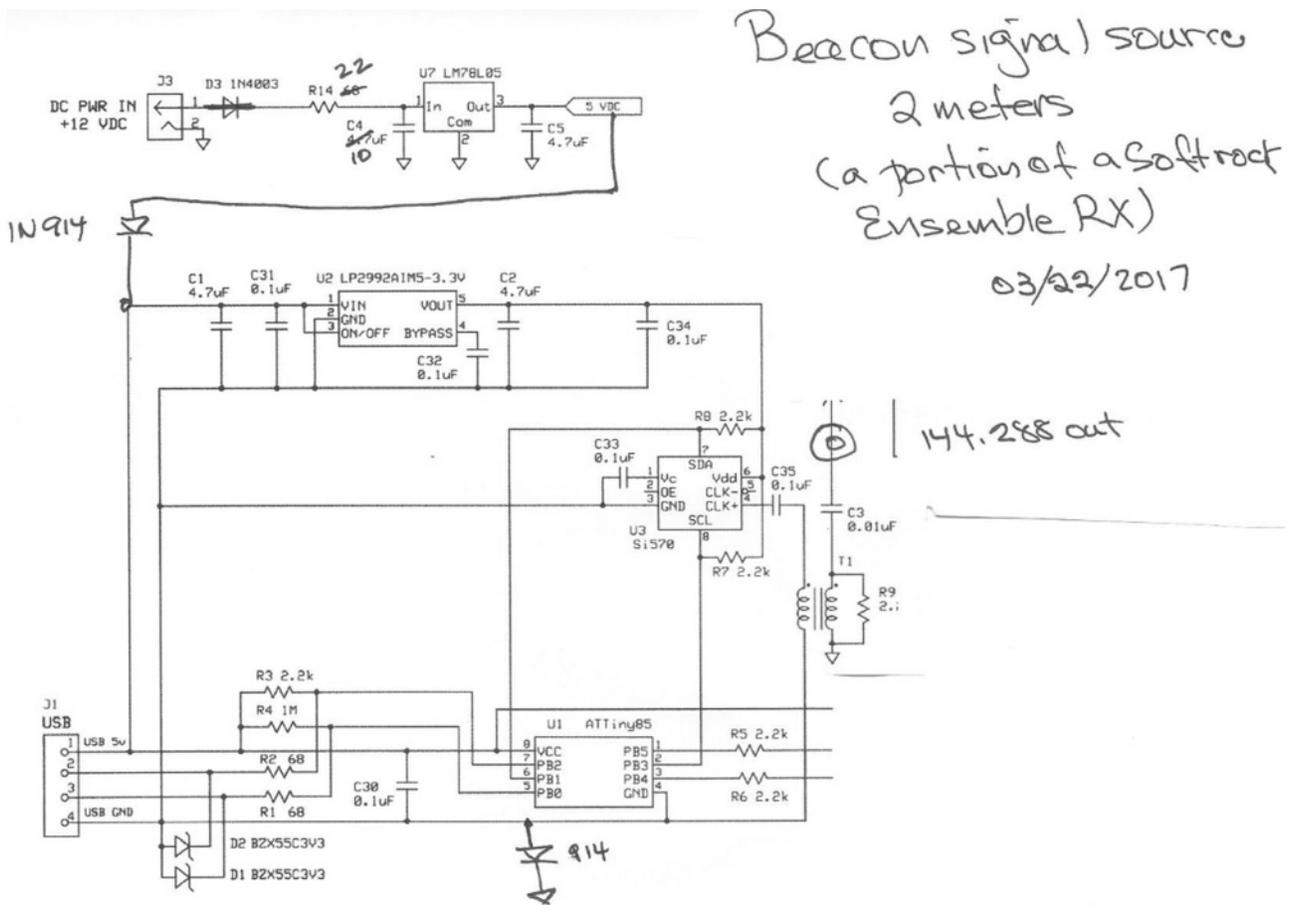


Figure 3. For 144 MHz (and below) beacons, merely employ the DDS portion of a Softrock RX Ensemble. Using the CFGSR utility, change the board’s startup frequency to your desired beacon frequency.

Tony Parks of Five Dash occasionally has returned Softrock kits with functioning DDS portions that he will sell at a much-reduced price. More than 12 dBm is available at the output of T1 of Figure 3, at any desired frequency from 10—160 MHz. With the change of a single line in one tab of PE0FKO’s CFGSR utility, the Softrock board can be set to start up at any frequency of your choosing. Once reprogrammed and USB power is removed, the board will wake up at the desired beacon frequency when powered up from 12 volts via the two added diodes shown in Figure 3. These boards also make excellent quiet LOs or lab frequency sources.

Unfortunately, the Softrock boards, with their CMOS Si570s, are unable to generate frequencies at UHF and SHF. Moreover, the firmware onboard the ATtiny85 controller locks out such programming choices. For these higher frequencies we must employ the Arduino/Daughter board of Figure 2.

Beacon sources at UHF and SHF

For these higher frequencies we replace the ATtiny85 controller of the Softrock with an Arduino (almost any Arduino will suffice) and employ a daughterboard populated by a faster LVPECL, LVDS, or CML version of the Si570. Again examine Figure 2. A typical daughter board (kit) is shown in Figure 4.

calibrated, there is a systematic procedure to exactly spot the desired final beacon frequency --- see Appendix.

Models of the Si570

The 3.3-volt Si570 comes in 36 flavors:

- Outputs: LVPECL, LVDS, CMOS, CML (A,B,C,D)
- Temperature stability: 50, 20, 7 ppm (A,B,C)
- Speeds: 10—1417*, 10—810, 10—280 MHz (A,B,C)

So for example, a Si570 with part number 570 CAC xxx-xxx-xx is a CMOS, with temperature stability 50 ppm, and speed 10—280 MHz, which is the Si570 included in Softrock kits. The remaining numerical xxx-xxx code is for start-up frequencies, the trailing letters xx denote packaging.

For UHF or SHF beacons we need the fast LVPECL, LVDS, or CML version with frequency range to 1417 MHz, but can live with the modest temperature coefficient of 50 or 20 ppm. That translates to the model number 570 {A,B,D} {A,B}A. My sketches for 220 MHz and up assume the use of the fast chip. A signal source using the fast chip and sketch could be used on any band by merely changing to the suitable N1, HS_DIV, and RFREQ in the preamble of the sketch.

If you happen to acquire a 570 *C*, there is one last complication: the six (one byte) registers of the Si570 holding the programming data are different for these 7 ppm chips. For the less expensive 50 or 20 ppm version the frequency programming registers are 7 – 12, but for the 7 ppm chip they are the registers 13 – 18. I resisted the temptation to automate this register change in the code. You will have to modify the register numbers in the sketches to use the 7 ppm chip.

Construction details

The si570 daughterboard, Arduino, and 5-V regulator can all be mounted together on a perfboard (see Figure 5), then enclosed with the keyer in a cast aluminum box --- see Figure 6.

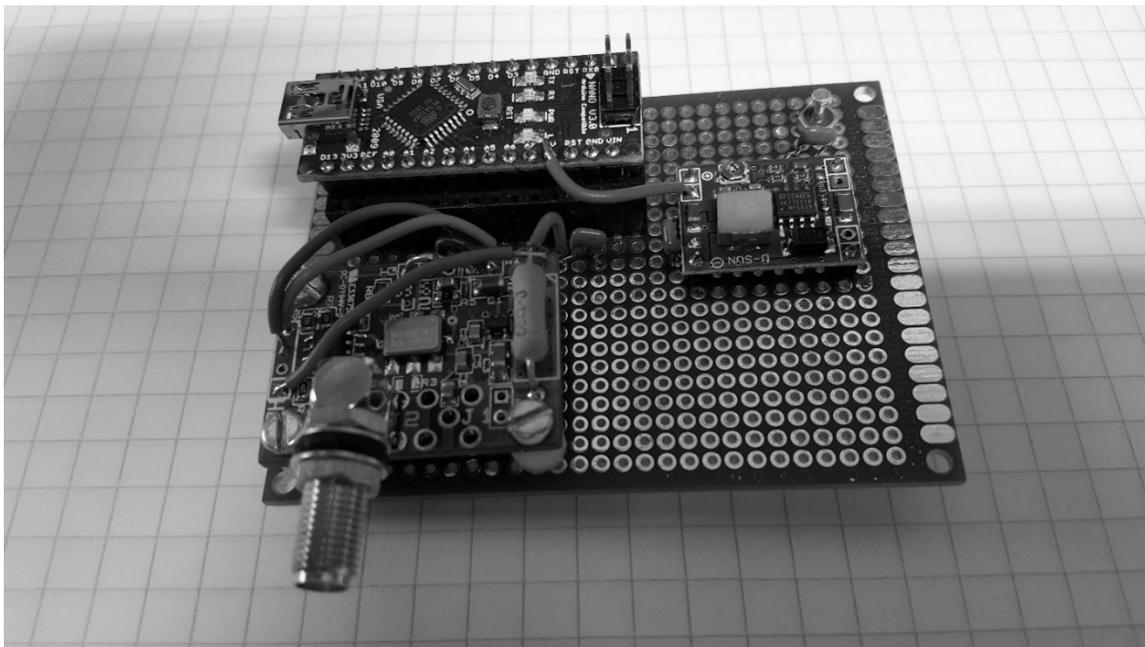


Figure 5. The si570 daughterboard, the Arduino, and 5-V regulator can be mounted together on a project board.

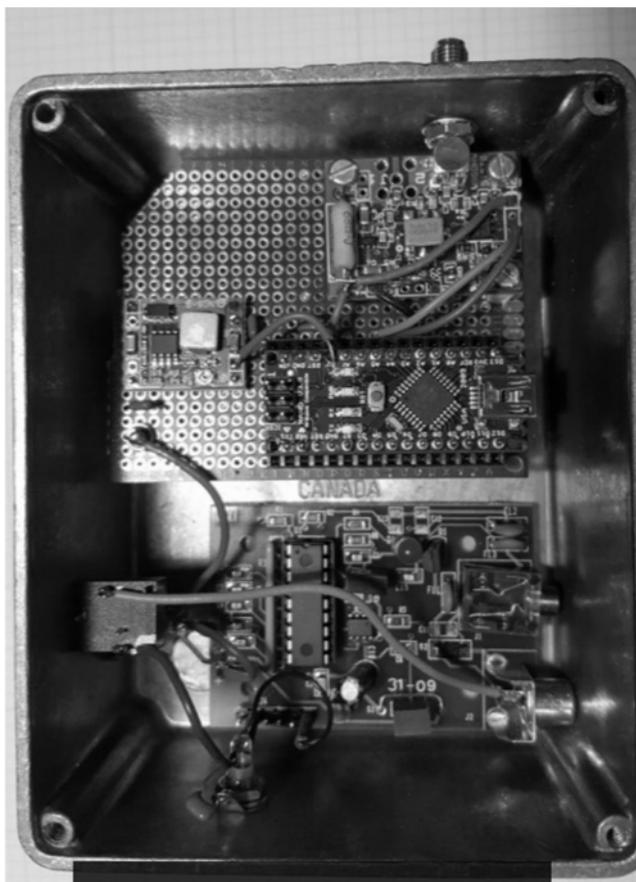


Figure 6. The signal source is enclosed with a KT-4 beacon keyer in a Hammond box with a 2.5 mm jack that brings in 12V, and a 3.5 mm stereo jack that brings in an amplifier sample voltage and the keying line. The RF is brought out via the daughterboard SMA and an RCA jack is used for external metering of the amplifier sample voltage.

The Downeast Microwave series of 30-watt amplifiers (2M30PA, 222PA, 7025PA, 3340PA, or 2330PA) make an ideal match for these signal sources, requiring only 8—9 dBm drive. These amplifiers include a PTT line out for the beacon keyer and an RF sampling line for a 1 mA meter. However, the CMOS Si570s crank out more than 12 dBm and overdrive the Downeast amplifiers, so that a series 5-dB pad will be required.

Concluding remarks

I urge everyone with a bit of tower space to do their civic duty by throwing up a quick beacon. The PAR omnis are perfect antennas for VHF and UHF beacons. However, 23 cm omni antennas are a problem. I would like to hear from you if you have solved this conundrum.

An unexpected annoyance arises from the use of switching 12-V power supplies to power these beacons: in synchrony with the CW keying, the power supply emits birdies on random HF frequencies; a linear power supply may be warranted.

I will be delighted to supply the Arduino sketch for any band you ask. Email me at w8mqw@arrl.net

Footnotes and suppliers

*10—945, 970—1134, 1213—1417.5 MHz

Unified Microsystems: <http://www.unifiedmicro.com/beacon.html>

<http://wb6dhw.com/Si570/Si570.html>

<https://www.downeastmicrowave.com/Default.asp>

APPENDIX

Refining the Arduino band sketches to place beacons exactly on frequency

This post is for readers who find cut-and-try steering to the desired final frequency objectionable. Here are the means of adjusting for the slight variations in the crystal frequency found in individual chips, so that you can directly program to your exact desired beacon frequency.

The frequency of the crystal onboard the Si570 is nominally XTAL = 114.285 MHz. Upload the sketch for your band into the Arduino, power up the signal source, and measure the actual output frequency. Then your actual onboard crystal frequency XTAL is found by multiplying the nominal crystal frequency by the ratio of the actual output over the expected. For example, using the 144 MHz sketch, if the actual output frequency is 144.274 instead of the expected 144.275, then your actual onboard crystal is oscillating lower at

$$114.285 * (144.274/144.275) = 114.2842079 \text{ MHz.}$$

Use this corrected value henceforth for your value of XTAL.

Next, choose your desired beacon frequency F_{out} . Then using the decimal values of N1 and HS_DIV in the sketch preamble, compute the decimal value of RFREQ by the rule

$$RFREQ = F_{out} * N1 * HS_DIV / XTAL.$$

Last, convert this decimal value into hexadecimal with the leading two hex digits representing the integral part, while the next seven hex digits represent the fractional part.

For example, in the 144 MHz sketch, N1 = 4 and HS_DIV = 9. Then supposing we want to place the beacon at 144.288, we see that

$$RFRFEQ = 144.288 * 4 * 9 / 114.2842079 = 45.45131909.$$

The integer part $45 = 32 + 13$, or hex 2D, which is written as 0x2D. The fractional part is transformed into a 7 place hexadecimal by using any online decimal-to-hex converter to transform the integral part of

$$0.45131909 * 2^{28} = 121,150,046, \text{ (rounding up)}$$

resulting in the hex value 0x7989A5E. The two hex numbers are concatenated to become the 9 digit

$$RFRFEQ = 0x2D7989A5E.$$

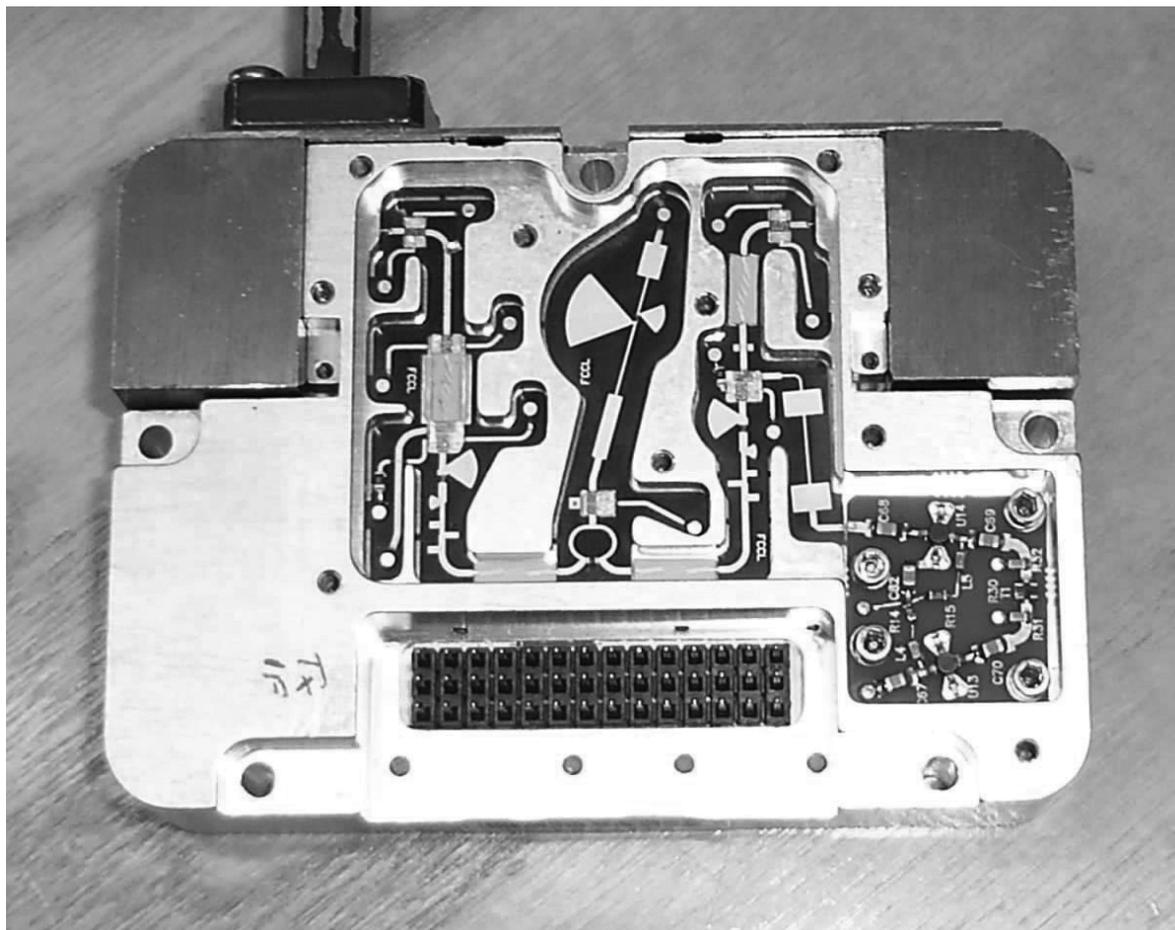
Replace the nominal RFREQ in the sketch with this new hex value and the oscillator will be spot on your chosen beacon frequency.

These computational details are explained quite well on page 19 of the Silicon Labs Si570/571 Application Note. As you might expect, when these details are incorporated into the sketch instead of done offline as above, the code begins to become overwhelming.

The Si570 is programmed by writing these three parameters N1, HS_DIV, and RFREQ into 6 registers in a rather idiosyncratic way --- examine my sketches and see the Application Notes. For small changes in frequency, only RFREQ need be changed. For large frequency jumps, the digitally controlled oscillator must first be frozen, then restuffed with a new triple (N1, HS_DIV, RFREQ), set, then unfrozen. All this is explained in some places clearly, others places not so clearly, in the Application Notes.

28 – 47GHz Amplifier

By Roger Ray G8CUB



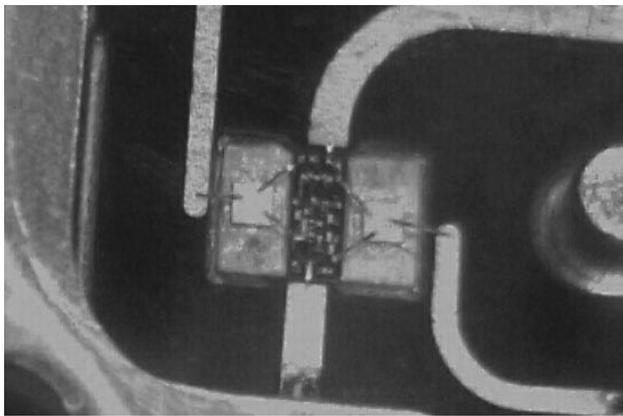
I was intrigued by a ‘milli-meter transceiver’ on ebay from ‘rf-buy2008’. Described as a ‘Signal Technology’ unit, but with no frequency information. From the dimension of the waveguide hole I guessed at 40GHz. What I was hoping for was >40GHz, what I got was 37.4GHz – bother! The picture shows the business part of the unit, with the duplexer and waveguide power coupler removed.

What looked potentially hopeful, was that it had a LNA amplifier on the receive side, and a driver and PA on the transmit side. Looking under the microscope, I identified the PA as an Agilent HMMC-5040. Looking further the LNA (marked 6561H) driver (marked 6742) & LO amplifiers (6561H) were UMS dies CHA? However asking UMS to identify them appeared akin to asking for the launch codes to France’s nuclear arsenal! Both mixers were Hititte devices, but again unidentified.

Firstly I decided to try the unit. Supplies were marked with voltages, except for the Tx amplifier supply, which turned out to be +5V to give around 4.5V on the PA supply. On the opposite side to the one shown was a VCO (V950ME01) which covers 4.3 – 4.65GHz. There are two X3 multipliers in the LO chain (back to back diodes), thus the LO was 38.7 – 41.85GHz. I connected the supplies, and applied 5mW of

2GHz to the IF in port. The output was +17dBm peaking (by adjusting VCO voltage) at 38GHz. Could I get anything out at 47GHz? I put a piece of foil on top of the filter, and applied 5GHz to the IF. I got at best -17dBm out at 47GHz. Not too good. Very little if I made the IF 5.76GHz, so I decided this was not going to work as a 47GHz transverter. In playing with the Tx filter, I broke the bond wire from the filter to the driver – double bother!

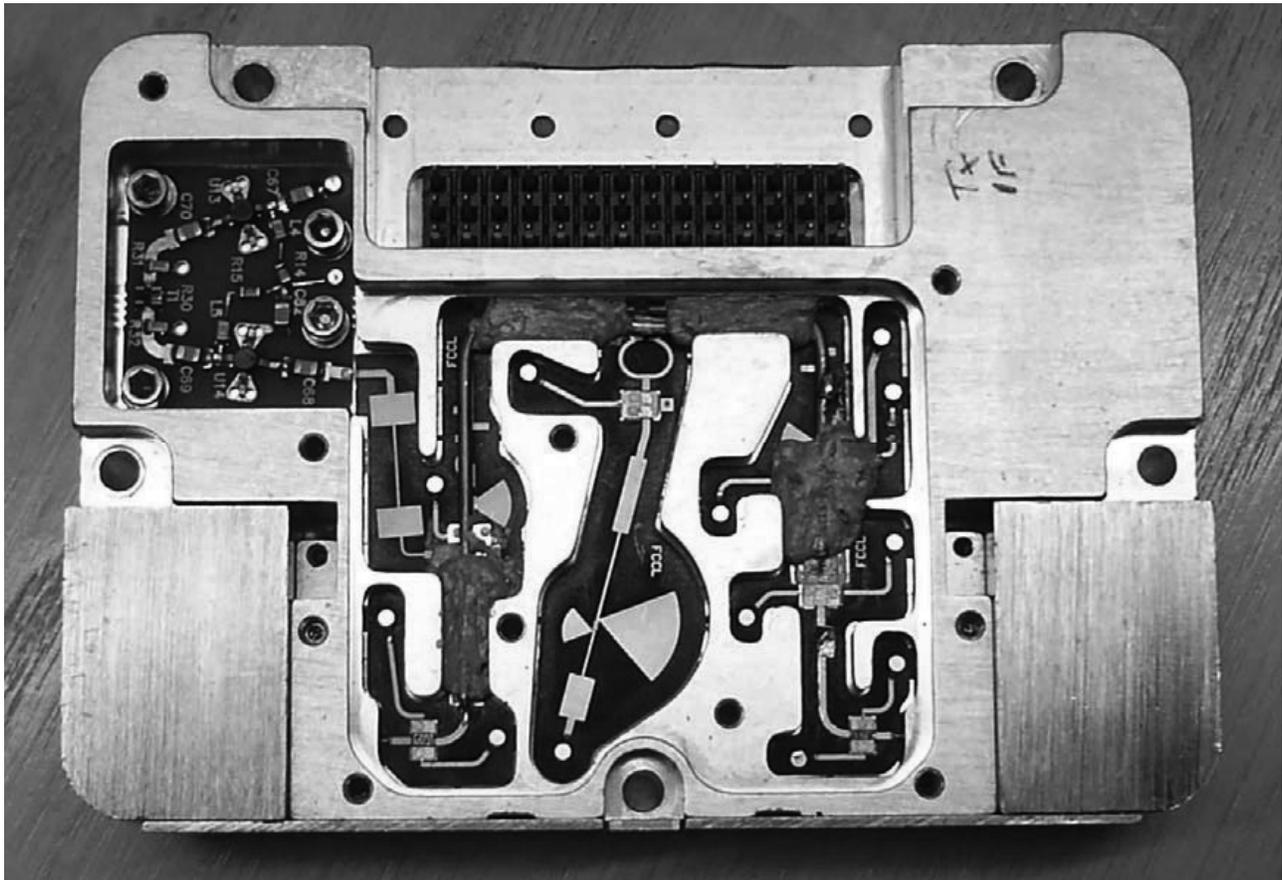
Could I make something of what was left. I first thought of trying to use the Tx and driver stages as an amplifier. However trying to find a way of coupling into the board was defeating me. Then I hit on the idea of using the wr-28 input and output ports. Would it be possible to connect the LNA through to the driver? The most likely outcome I guessed if it worked at all, was that it would oscillate. Nothing ventured nothing gained etc, I gave it a try.



Agilent PA



Driver

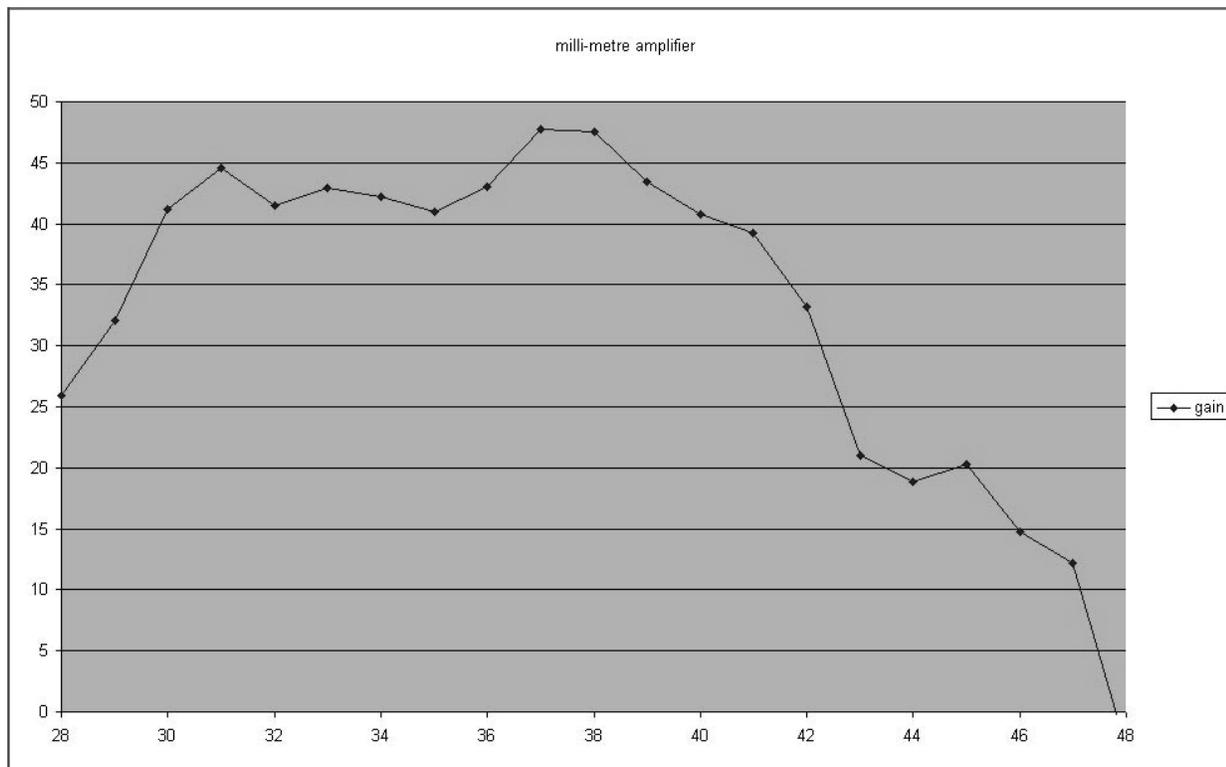


I used miniature coax that was to hand, as it was likely to propagate the high frequency and give some attenuation between LNA and driver. I soldered the inner on to the track at the LNA end, and used liberal amounts of silver epoxy to connect the outer. This I found I needed to cure, so I put it in the oven at 100 deg.C for an hour, which seemed to do the trick. At the Tx filter end, I bent the inner down and used silver paint to make the connection.

I disconnected the positive and negative supplies to the parts that I had now shorted, and applied power. Firstly I very pleased to find that I had gain, and secondly that it was stable! I did a sweep from 40 – 48GHz and found very healthy gain at 40GHz - 41dB! but very little at 47GHz. The amplifier was at its best with the lid in place.



Test set-up shown measurement 40 - 48GHz, 47.088GHz on screen. This used a Wiltron 40 -60GHz generator and a Flann attenuator and transition WR-19 to WR-28. On the output side a transition to WR-22 and WR-22 attenuator fixed at 40dB and a 33-50GHz HP harmonic mixer with HP 8563E analyser. I later checked the 28 - 40GHz response using a Spacek doubler on my sig gen, and an MI power meter.



The response shows gain in dB on the vertical scale, against frequency along the bottom in GHz.

Gain was a massive 47dB at 38GHz, 40dB at 40GHz, but only around 10dB at 47GHz. 1dB compression was 21dBm at 40GHz, around 18dBm at 37GHz, but only 8dBm at 47GHz.

To get the best performance at 47GHz, I added a small tab on the line between driver & PA, and turned the current up on the amplifiers a bit. At best I got 11.2dB at 47.088GHz, but could not quite repeat that later.

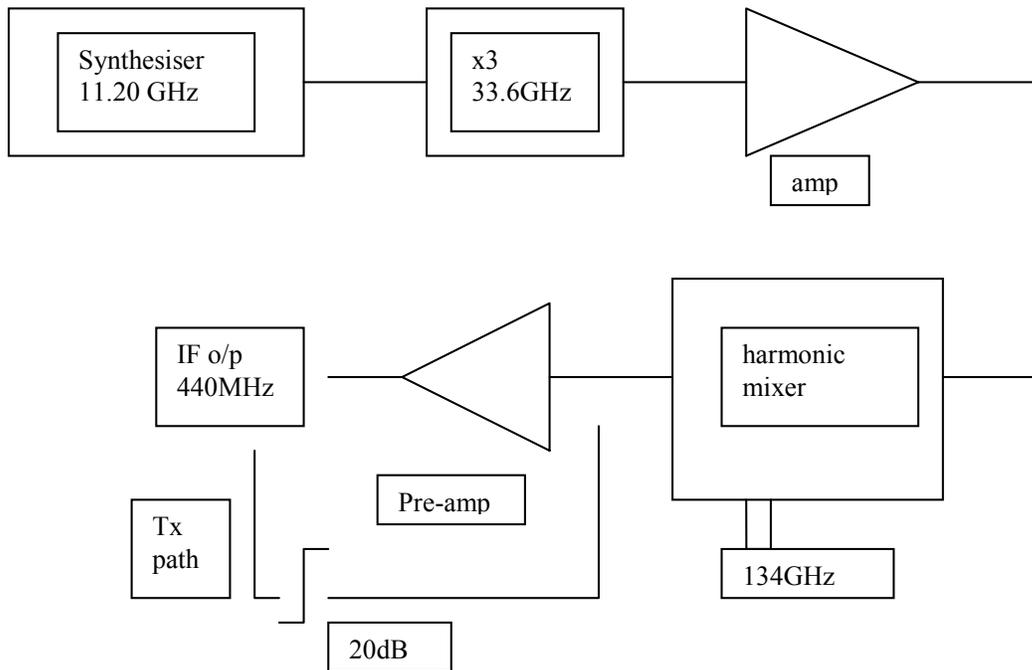
There may be more to be had by adding tuning screws to the waveguide ports, and more careful tabbing. Also I think that if I had ground off the unwanted filter elements it might of helped – though I could easily have broken another bond wire, in doing so.

So was it worth it – as a 47GHz amplifier probably not. As a wideband amplifier to drive a mixer X4 for 134GHz hopefully! For this application it could well be easier to use the internal last X3 multiplier in the LO chain. In this way a coax input could be used after the LO splitter and the following filter. The Tx mixer would need to be shorted out, and possibly the final Tx filter shifted in frequency.

At the time of writing units are still available on ebay.

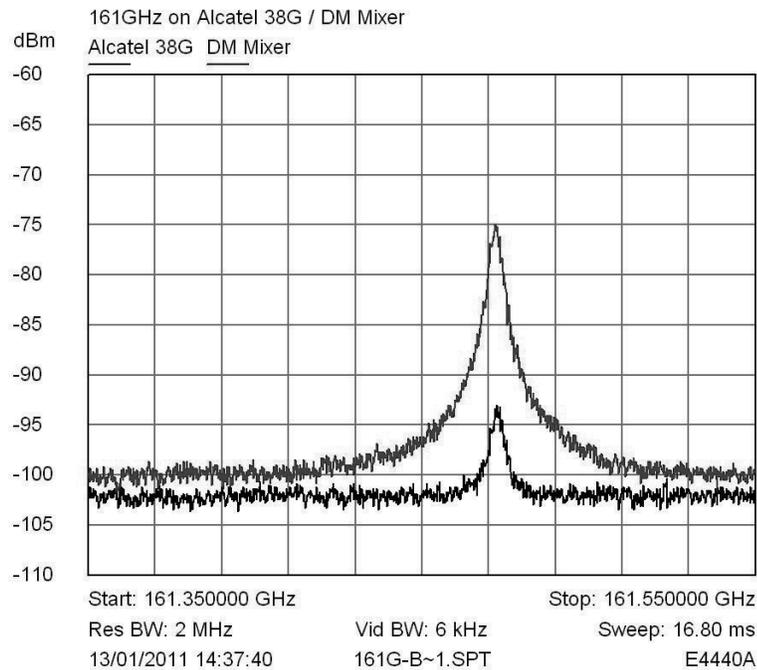
134GHz Transverter

Roger Ray G8CUB



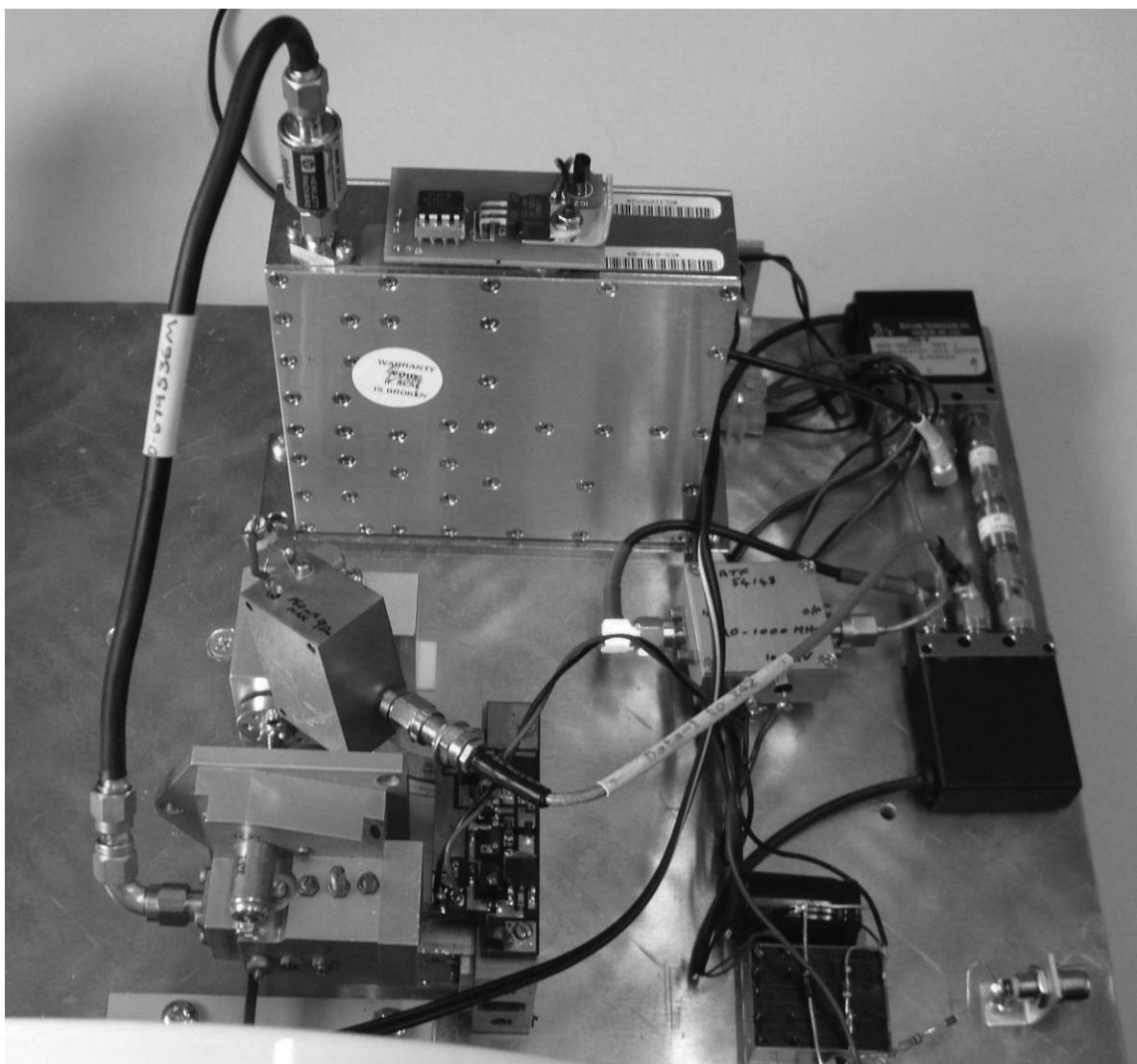
Simplified 134GHz transverter block diagram

The idea for a 134GHz system started about 18 months ago. Initially I intended to use an Impatt diode source. This proved to be very noisy, and hence requiring a large bandwidth.



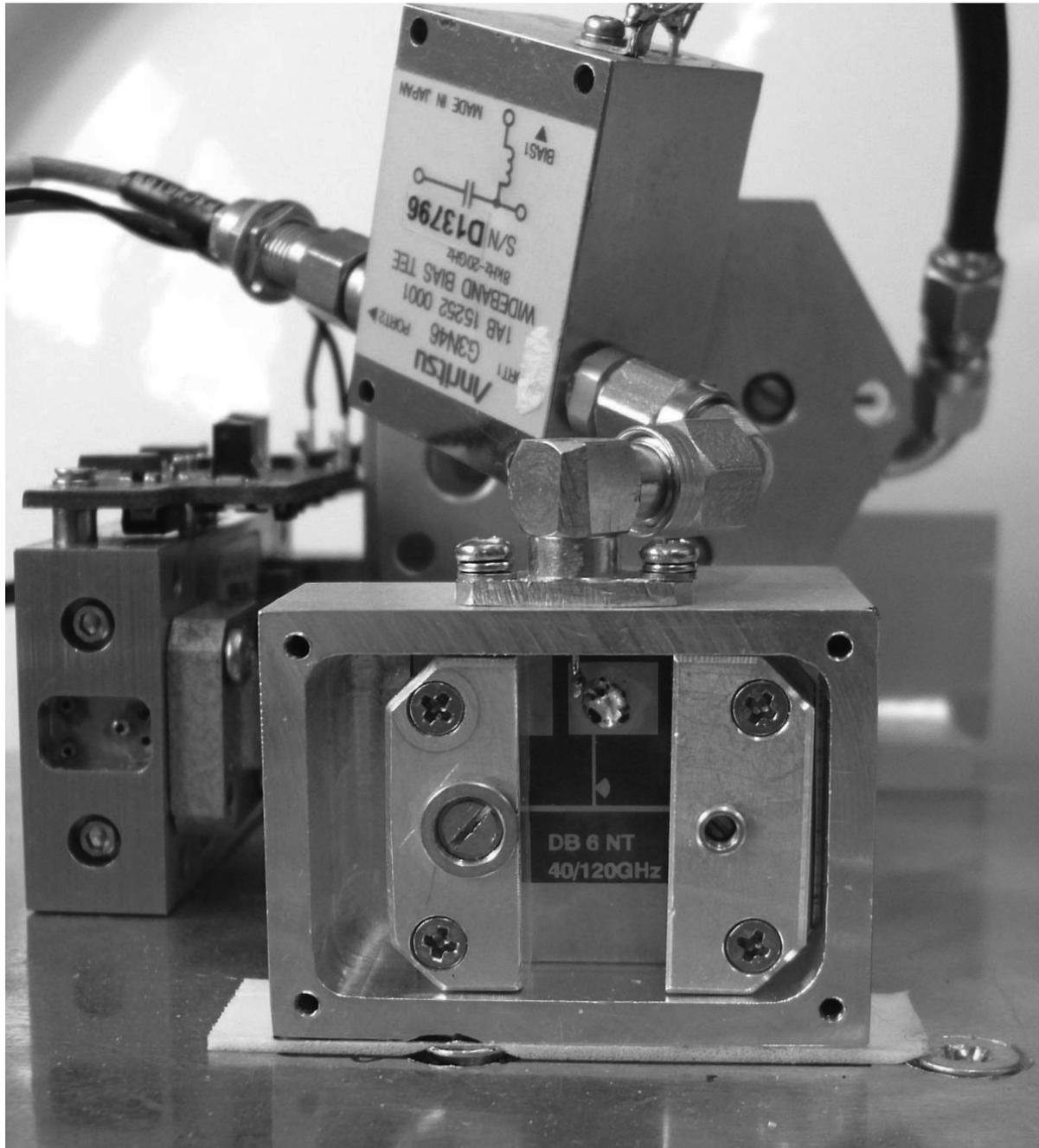
The plot shows the oscillator on 161GHz. The higher (blue) trace was measured with a commercial mixer (110-170GHz), the lower used a 'test module' from an Alcatel 38GHz 'white box'. This shows that even at 160GHz the test modules work as a low cost mixer! The signal bandwidth is much greater than the spectrum analyzer 2MHz bandwidth, so the true power is greater than the plot.

Having rejected the Impatt diode idea, I looked for a way of using an Elcom synthesizer to drive a harmonic mixer. The Elcom lowest frequency is 11.200GHz, with 3.333MHz steps. Doing the maths, showed that $11.206666\text{GHz} \times 12$ gave 134.480GHz. It would be also possible to produce a TX signal at 134.920GHz (11.243333×12), giving a 440MHz IF. With a 134.48GHz LO, it would be also possible to receive the existing sub-band of 134.928 – 134.930, and transmit using a modified FT817 (448-450MHz IF).
[Also a 135.360GHz LO (11.28×12) could be used to give an IF of 432 -430MHz]



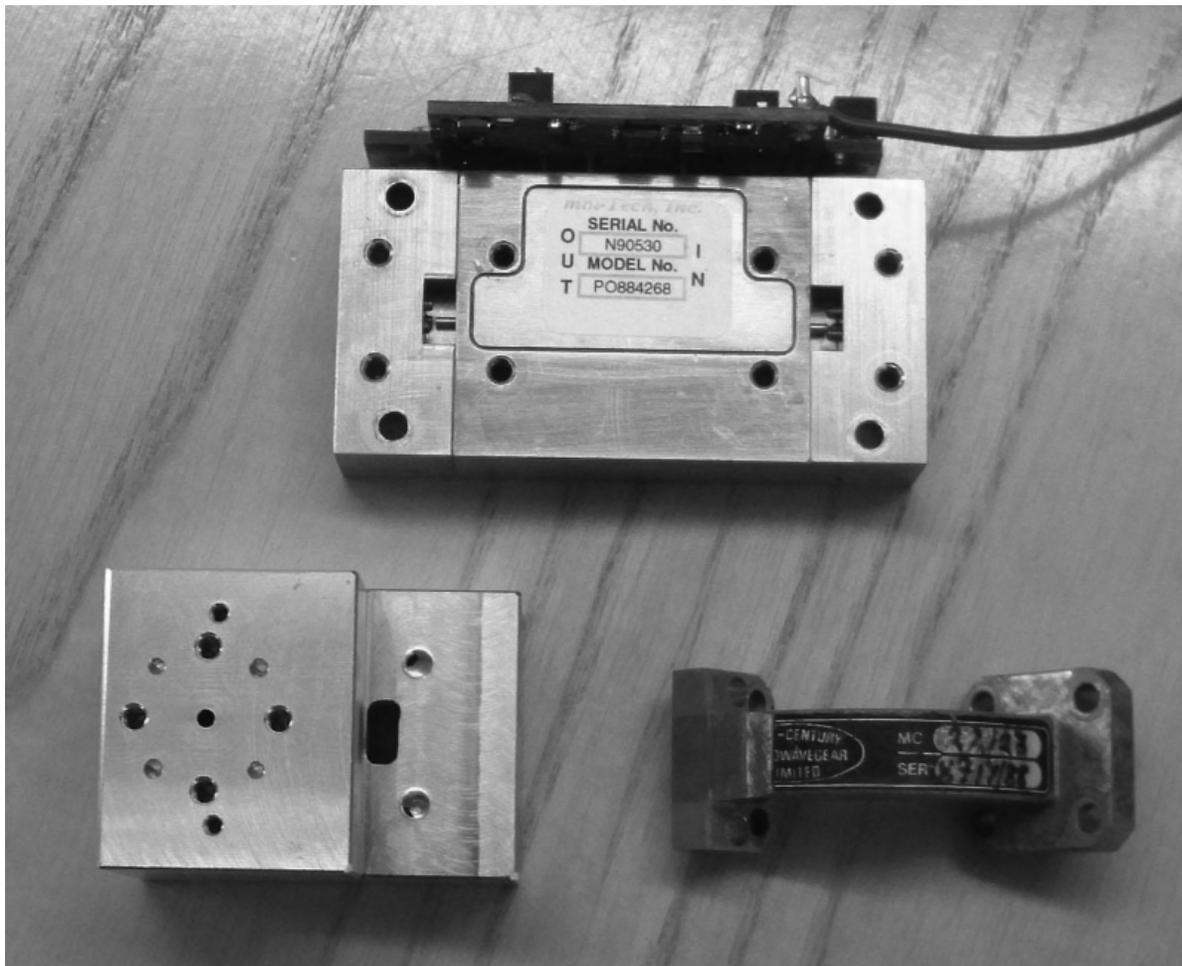
The synthesiser is at the top of the picture, with it's PIC loader board on top. Firmware is from Dave G4FRE, and is available on his website. In the foreground is a X3 tuned multiplier, which drives a mm-tech 23 -36GHz WR-28 amplifier. It was the appearance

of these amplifiers on eBay for 20 Euros, that sparked of this version of the project. On the right are the TX/RX relays and UHF pre-amplifier (ATF54143).



The harmonic mixer is a DB6NT pcb No.47. The case used is a 122GHz housing from Philipp DL2AM.

The only difficult bit here, is that the case is designed to use a DML type 38GHz multiplier. To use a waveguide input, I modified a WR-28 bend by machining off part of the flange. This can be seen in the next photo, together with the mixer block, and mm-tech amplifier. After machining (or sawing/filing), it is then drilled to match the taped holes in the block.



The only part to mount on the pcb is the diode. Here there is a choice. Starting at under 7 Euros is an MAE1317, up to an HSCH-9161 at 45 Euros. The potentially difficult part is mounting the diode – they are VERY small. There are various schools of thought, as to the best mounting method. I used 2 part silver loaded epoxy, to attach the diode. A good microscope, is a must, preferably binocular. The adhesive is applied to the board, and the diode put in place using an alcohol wetted cocktail stick. The board is then cured at 100 deg.C for 2 hours. I then fitted the board into the block, which is held down by the two tuning element blocks. Articles in Dubus suggest that the board should be glued in place with silver epoxy. I have not done that, as I prefer the freedom of being able to change the board. The diode is self biased, using a 1k pot to ground. I used a bias-T to separate RF and DC. However a simple choke and capacitor would have sufficed.

The amplifier was hard into saturation with an input of -2dBm. After tuning the 3 screws in it's output waveguide, I obtained an output of +15dBm at 33.6GHz. With the pot set to mid-range the voltage across it was 1.33V. This is after adjusting the tuning elements for maximum voltage.

Do not do as I did the first time, by adjusting in the element over the diode and crushing it! I suggest setting the screw just over the diode and tuning out. This screw is finally set for best received signal, or maximum output.

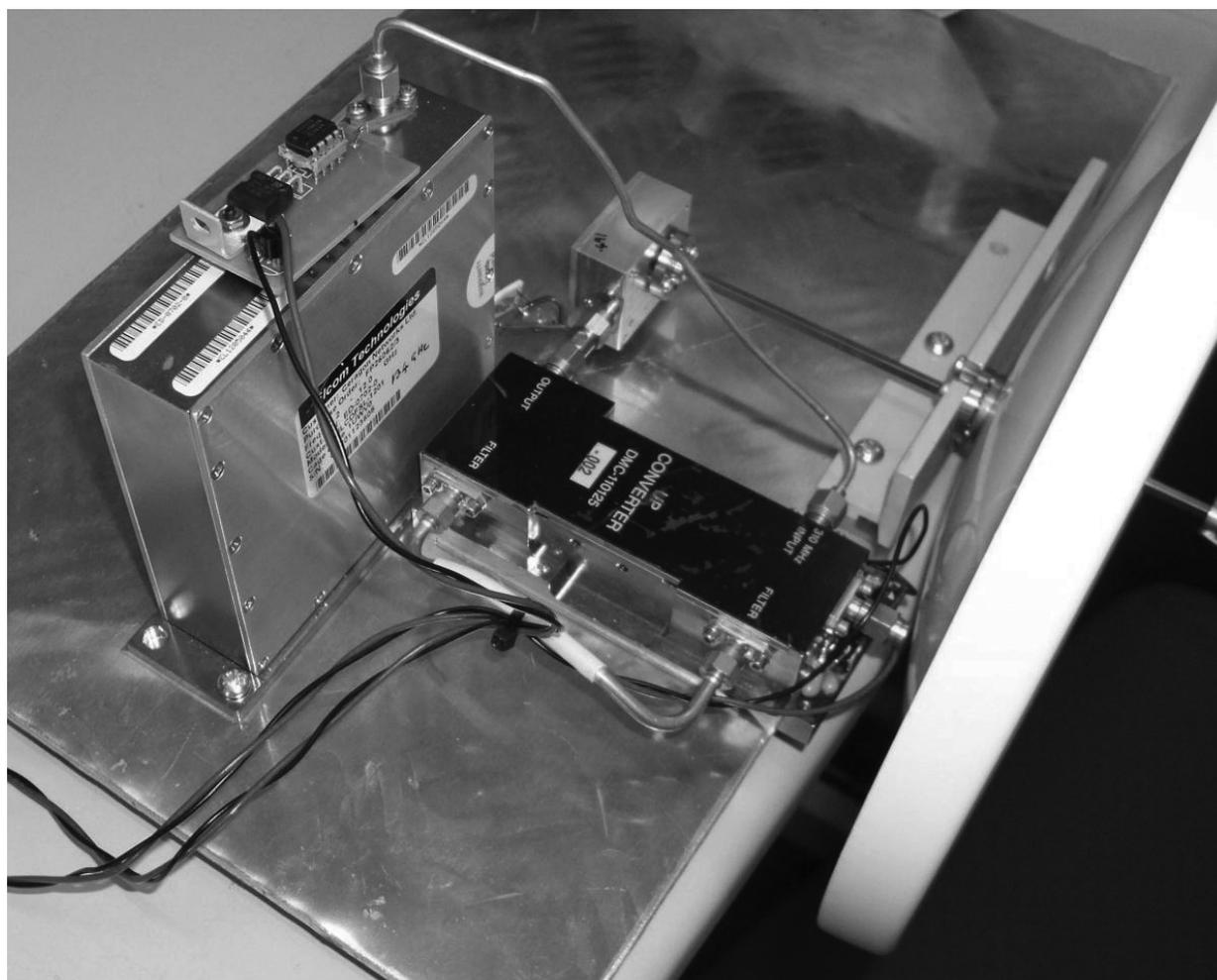
A pre-amplifier is necessary. I used an ATF-54143. A low cost pcb is available from RF Bay on eBay.

The selectivity of the mixer, is defined by the hole size in the block under the diode. This is used as a circular waveguide. The 122Ghz block came with a 1.85mm hole. This works fine at 134Ghz, but does not remove the unwanted signal at 101GHz. A possibility here would be to drill out the block and fit 1/16 id brass tube. I have now done this on a second unit, using 1/16" aluminum tube that was actually 1.65mm id. It works well and kills the 101GHz signal totally.

The second transverter I built by upgrading the 'test source' described below. Here I used the low cost MAE1317 diode. This unit had 6dB more Tx output than my first one, although the receive performance appeared lower.

My test source used another Elcom synthesiser driving a DML up-converter, used as a doubler amplifier. The mixer with RF on its IF port, and is then connected to the amplifier section. The output was +19dBm at 22.4GHz. This then drove a 76GHz block using an MA46H146 diode. A small amount of RF at 134GHz is produced, and is selected by the WR-7 waveguide, connected to a Procom dish.

This worked surprisingly well, with the test signal being around 60dB above noise at 8 metres away.



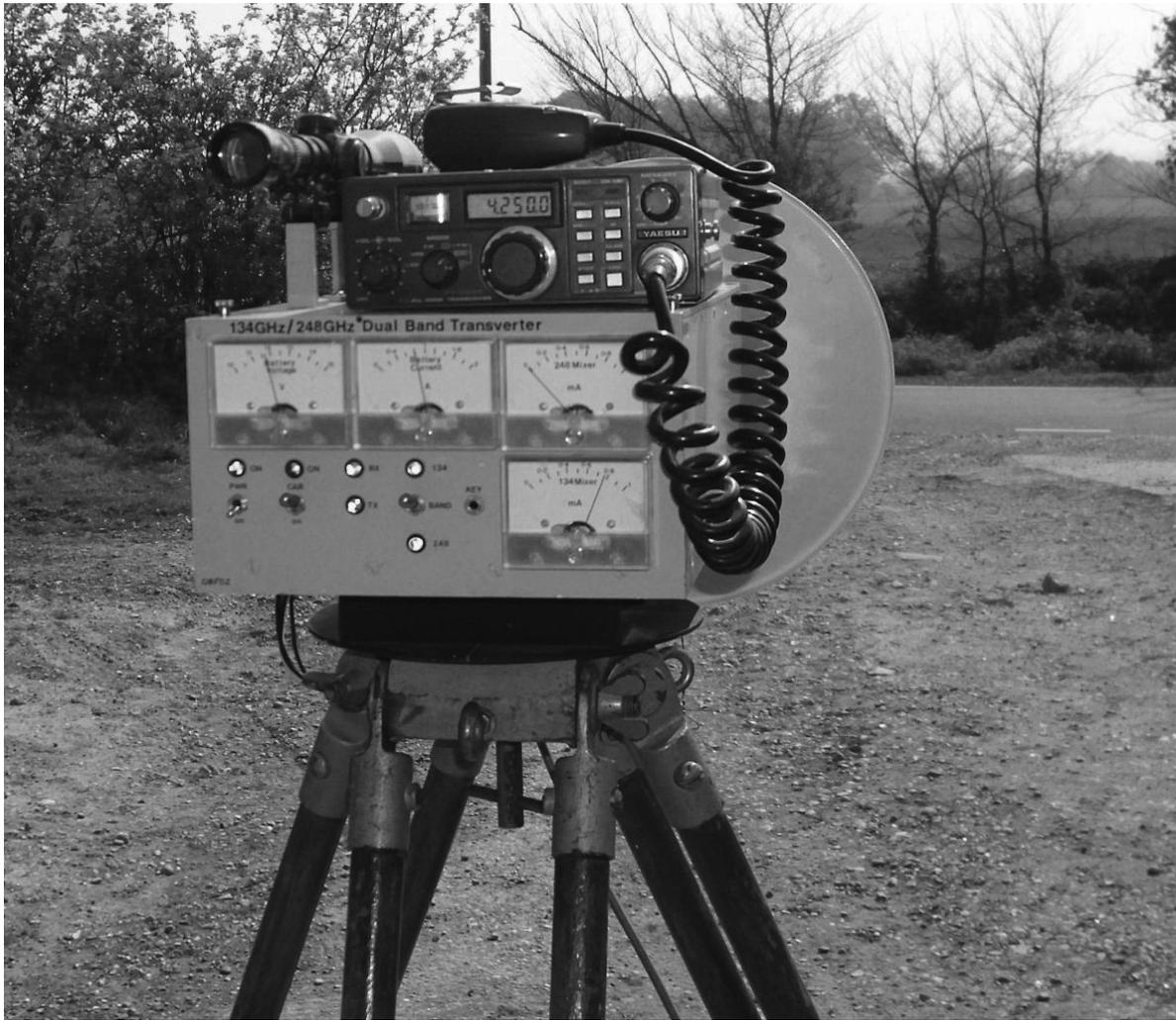
It is intended to later convert this test source to another 134Ghz transverter- now done!

The first on-air tests were with Chris G0FDZ. After an initial false start, where it was found that Chris's transverter had a 1.4mm circular waveguide hole, that was cutting off at 134Ghz! It was drilled out to 1.6mm, and we were away. Well, we could copy each other at a couple of meters.

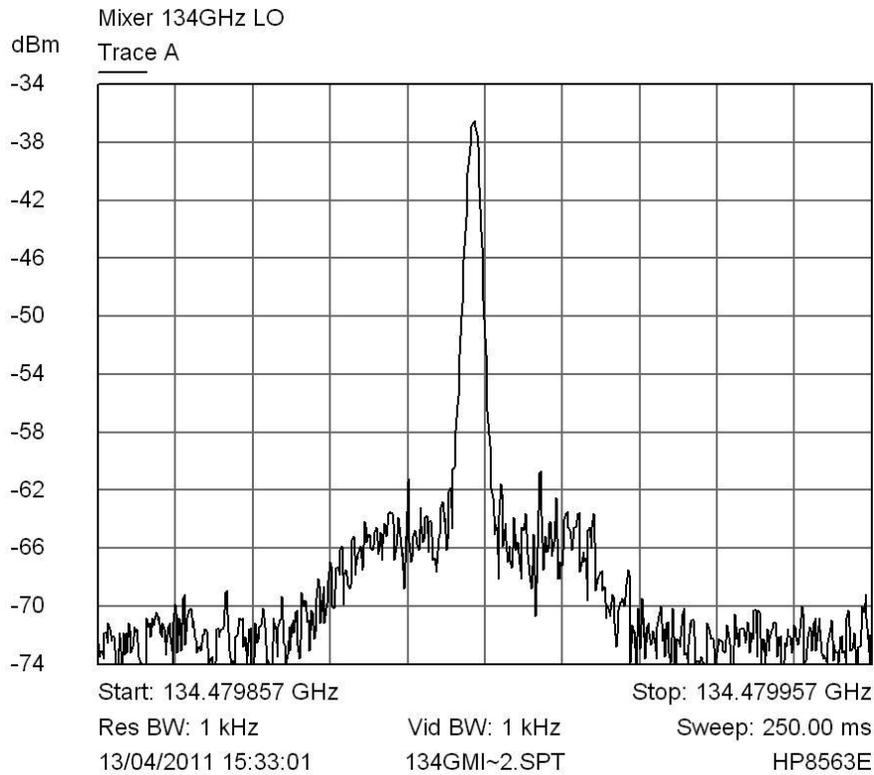
We struggled to find a clear test range, but eventually tried over a 500m distance. Alignment proved a challenge. The Procom 44dB dishes are very sharp, with a stated beamwidth of 0.8 degrees. Reports were exchanged on SSB to complete a QSO.



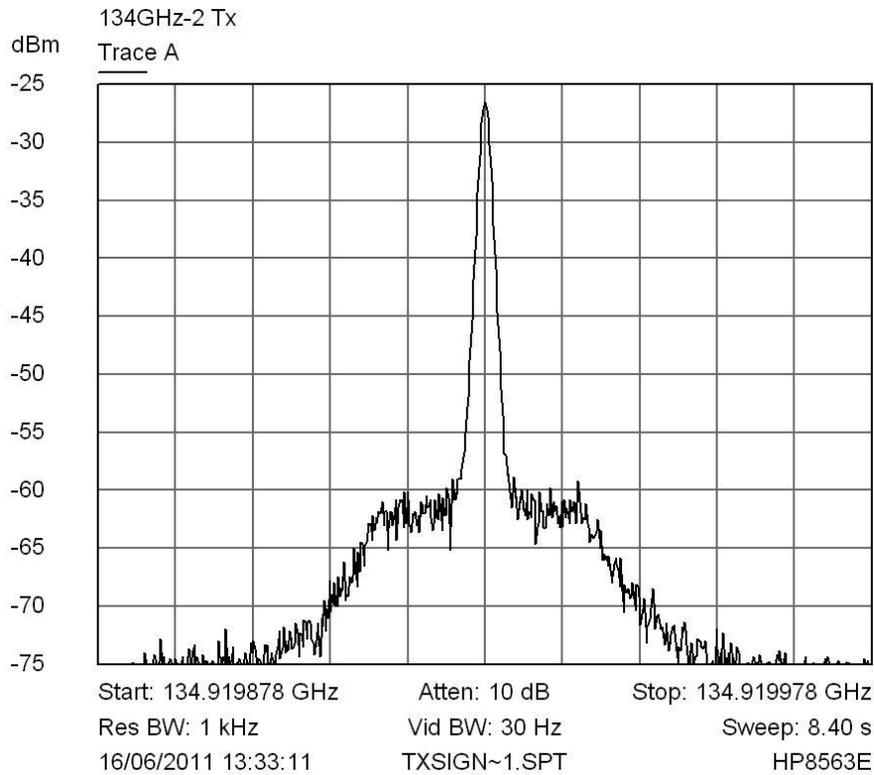
The transverter in action – yes I forgot about having to plug the microphone into the FT817!



The very well constructed G0FDZ transverter, at the 'farm' end of the link. Chris uses a 144MHz IF.



The spectrum plot of the LO signal at 134.480GHz. This using the internal 10MHz reference in the synthesiser. I am staggered how clean and stable the signal is! However since that first QSO, I found that the signal was not as clean when on battery. I was using a series diode in the supply. When the supply dropped below 11.4V the LO moved frequency and spurs were produced. Something I have never noticed at fundamental! It is fine on 12.0V.

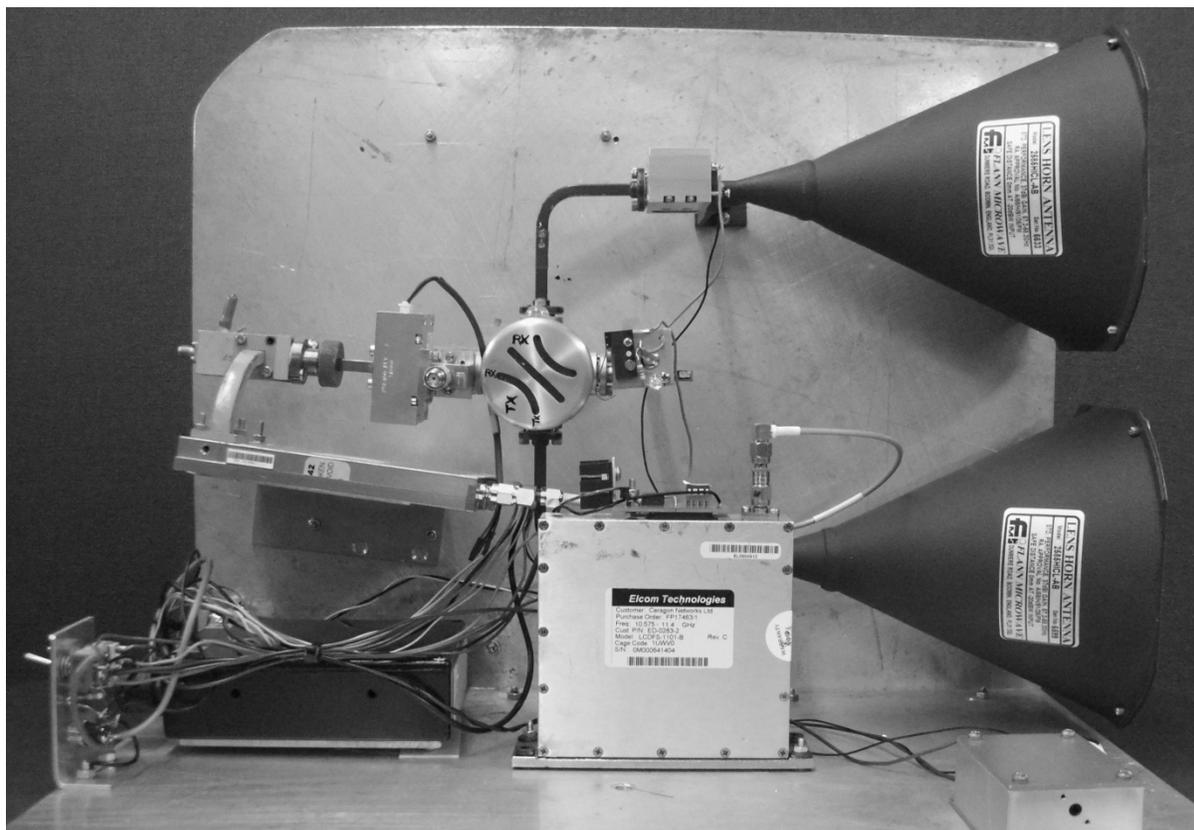


This is the Tx output spectrum of the second transverter. The true output power is greater than that shown as the mixer is calibrated with a 13GHz LO, and the spectrum analyser LO output is being used to drive it (around 4GHz). This is using the MAE1317 diode and an IF input of +11dBm. LO drive at 33.6GHz is just under +17dBm.

A 76GHz Transverter

Doing things differently!

By Roger Ray G8CUB



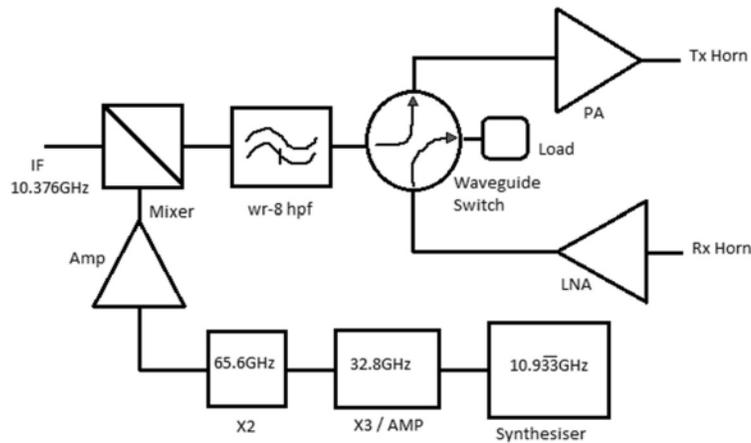
The concept for this 76GHz transverter came about following tests in August 2012. At Ditchling beacon at that time, only John G4EAT could copy signals from Ian G8KQW/P near Ventnor IOW. The difference being that John's transverter was using a fundamental mixer. Chris G0FDZ and myself were using DB6NT type harmonic mixers. These probably had around 16dB NF compared to 8dB or so for the fundamental mixer. John was also using a surplus 58GHz Flann horn, which appeared to work just as well as my 77GHz version.

So to design a new transverter.... Somehow component parts appeared pushing me in direction I eventually took. Harold G3UYM returned my small 10GHz transverter that he had been using. Ebay turned up a pair of 58GHz horns and separately a 60GHz receive system. The later had a plastic dish, a filter, amplifier and harmonic mixer. The amplifier was an Arcom 60N00 which showed around 18dB gain 46 – 70GHz. Not immediately useful, until I hit on the idea of using a 10GHz IF. The advantage of using a high IF is that it would allow image rejection with a length of WR-10 waveguide (cut-off 59GHz).

Looking at what else I had available. I had a Farran fundamental mixer (probably 0-4GHz IF). The Arcom amp and the 2 horns. I also had a WR-15 waveguide switch bought at Martlesham a couple of

years ago, and of course the 10GHz transverter. Also I wanted to start with an Elcom synthesiser, as they were proving to be so good.

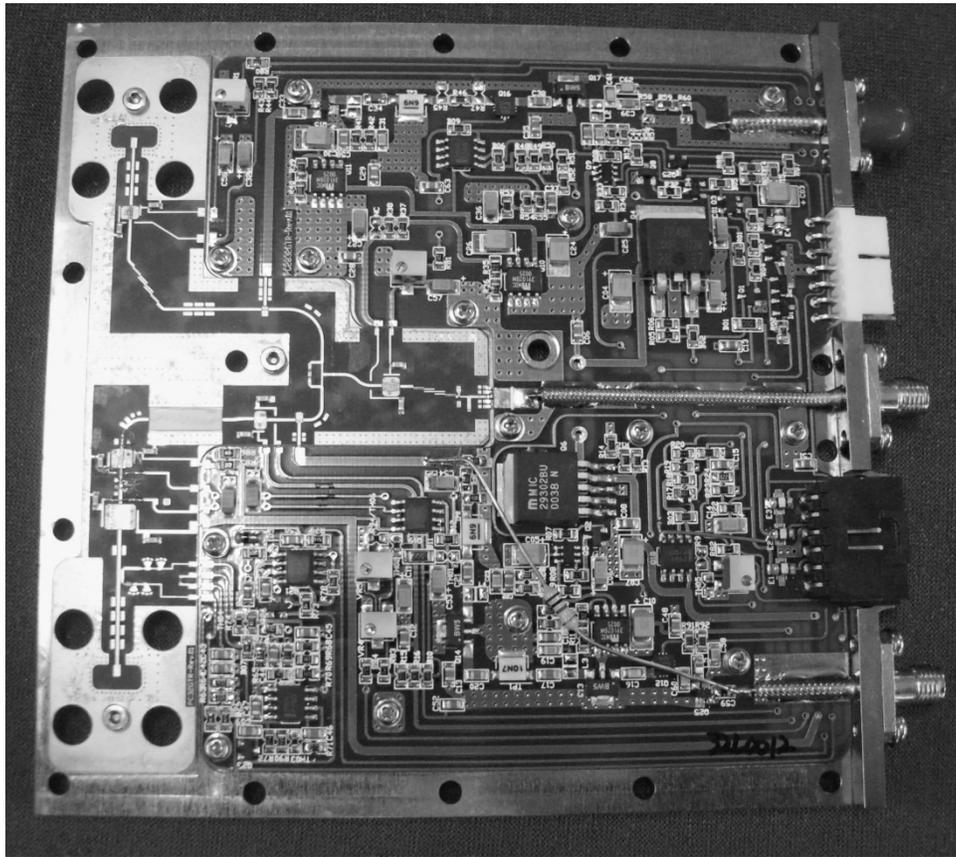
A quick calculation showed that if I used a 65.6GHz LO, I would have an IF of 10.376GHz. This with the existing transverter would give a final IF of 440-442MHz – not a problem with an FT817 (tx delimited).



Transverter block diagram. Waveguide switch shown in TX position

Unfortunately the Elcom synths that I had started at 11GHz, and I wanted $65.6/6 = 10.933333\text{GHz}$. There were however back on Ebay some Elcom units starting at 10.575GHz. A check via Doug VK4OE and the Australian group, responded that the programming was the same as for other Elcom synths – 3.333MHz steps. By a stroke of luck, $65.6/2$ fell within the range of the Broadern module that I had been using on 134Ghz. Just by adding one resistor the acted as a x3 multiplier / amplifier giving up to 0.5W at 32.8GHz! The trick here I had previously found, was to run both positive supplies at 5V and use a DC bias (via the added resistor) to make the mixer work as an attenuator. As long as the input to the LO port is kept below +3dBm, varying the DC voltage on the mixer gives a controllable output.

To get to 65.6GHz I tried a WR-28 mixer used as a doubler. I have found that generally if you drive the RF port of a mixer, they work well as a doubler, extracting the output from the LO port. Sometimes it is necessary to bias or ‘tune’ the IF port for maximum output. The only mixer I had was a fixed tuned 38GHz unit. A one off effort to move the tuning blocks, produced an output of -4dBm at 65.6GHz. The output was via a piece of WR-15 waveguide. This had to be carefully positioned and angled to get the maximum output. This is the reason that the Broadern module is mounted at the odd angle in the transverter.



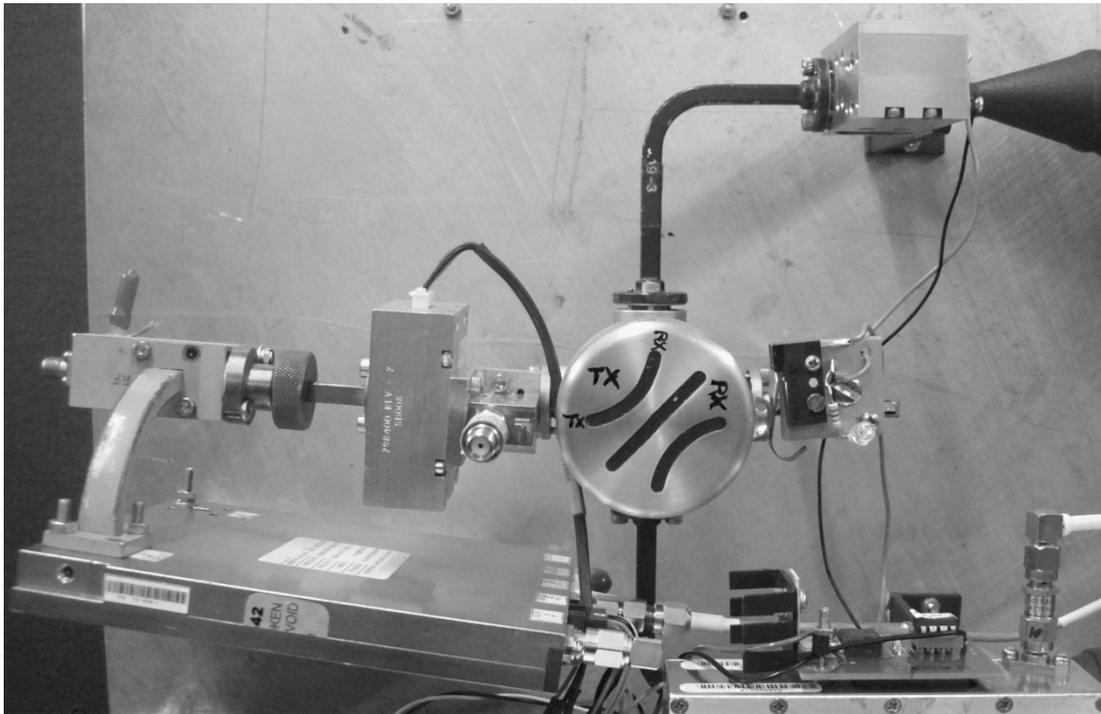
Broadern ED-0296-2 module – resistor added

The output from this mixer used as doubler went via the Arcom amp to the mixer. The amp produced around +11dBm maximum. This was ideal for the fundamental mixer. Input to the mixer was initially a length of WR-10 to reject the image. However when later the PA was added, I found the LO level too high. The solution was to change to WR-8 waveguide (cut-off 73.8GHz). This nicely removed image & LO, to <-60dBC.

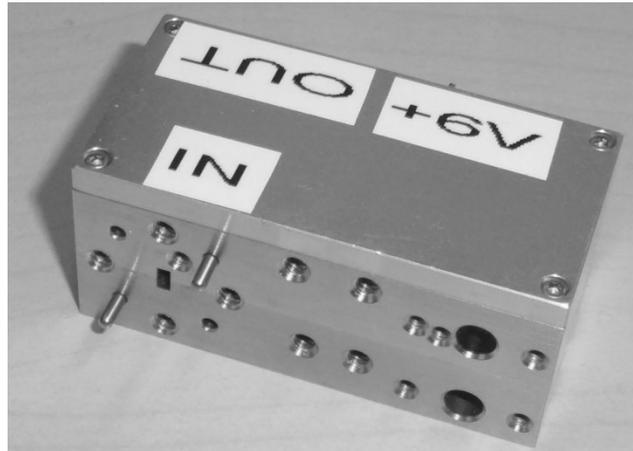
Initially the transverter was used like this, going straight through the WG switch to a single horn. On TX 1dB compression occurred at -3dBm output, with a drive of +12dBm at 10GHz.

Just after this the pre-amplifier from Tom WA1MBA arrived, the culmination of a 7 year project, to produce a batch of pre-amps. This had over 30dB gain and a measured noise figure of 5.0dB. I have now made my own noise figure measurement on the complete system, using a borrowed noise source – the average of 3 measurements was 6.1dB. This equates well with Tom's measured noise figure of the pre-amp.

The same week the PA arrived from the States. Claimed output was 100mW with +23dB gain.



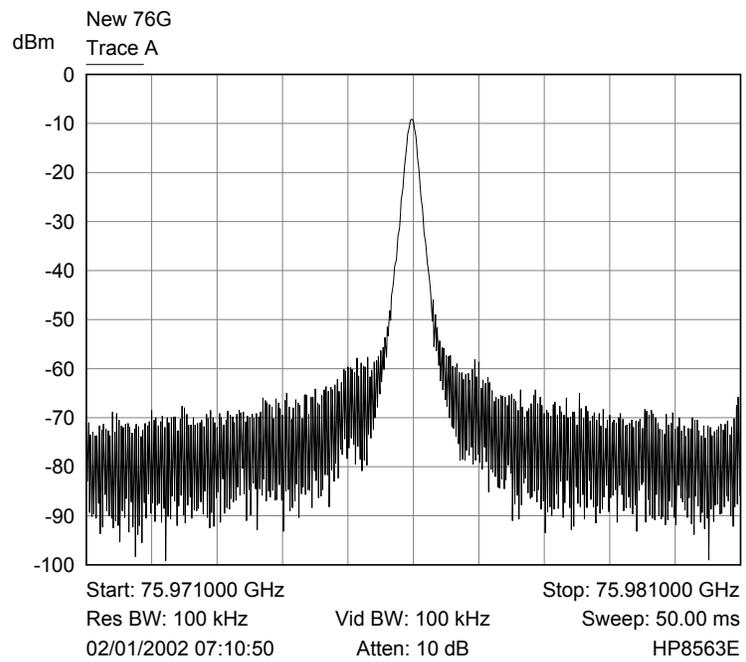
Waveguide switch in TX position



The 100mW PA – single +6V supply

Adding the amplifiers required two waveguide WR-12 bends, that I didn't have. These were eventually located in the States surplus, and duly arrived. What I had forgotten was the orientation of the switch and horns. To get horizontal polarisation, I had to mount the whole assembly vertically, as can be seen in the pictures.

With the 6 position WG switch there was now a problem. Two horns connected via over 50dB of amplification at 76GHz, in an intermediate switch position! To overcome this I added a plastic 'foot' to the switch knob. Only allowing the TX amp to be on when the switch was in the right position.



Tx output of +19dBm

Performance of the transverter has been excellent. Reports were 59/59 FM with G8ACE, and 59 / 599 SSB/CW with G8KQW from Firle beacon over a 94km path.



In Operation at Firle beacon with 10GHz transverter and FT817 fitted.



How the transverter looks in 2019, with 30cm dish, and two WG switches.

The space on the right is for the 10GHz (IF) Transverter.

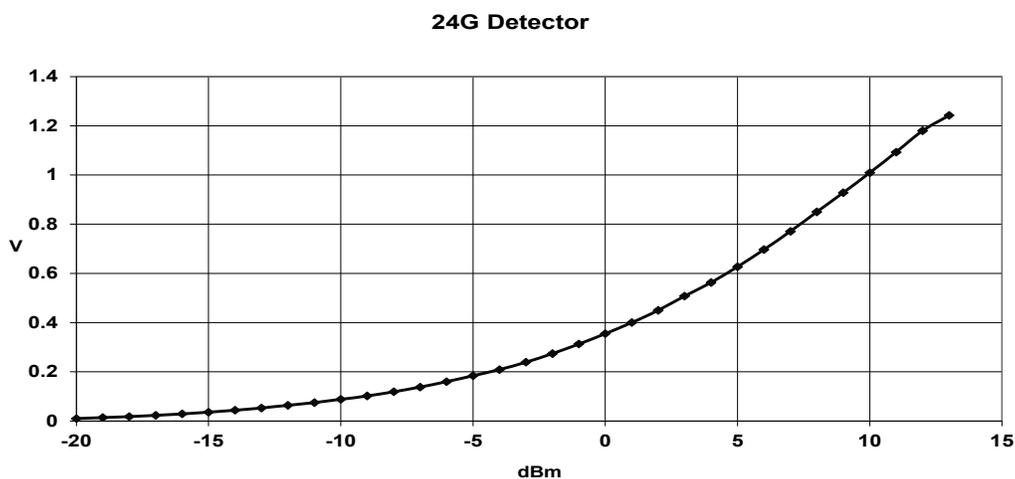


Alcatel 'Test Module' by Roger Ray G8CUB

It wasn't until I was looking for something to help me tune up the Pasolink 50 transceiver on 47GHz, did I discover this little module. It had been sitting discarded, in my junk box, since building a 24GHz Alcatel based transverter. It is originally situated on the coupler in the Alcatel ODU. It is used as both a detector, and a method of providing 'loop round' normally being fed from a dedicated synthesiser at 1008MHz (difference frequency between Tx and Rx).

In one go it answered my requirements for an RF detector, harmonic multiplier, and most surprisingly a harmonic mixer. Its use at 24GHz, I later found had been given in an article by G4JNT #. Operation as a detector / multiplier at 47GHz and 76GHz was less obvious, and performance as a harmonic mixer a revelation

Use as a Detector



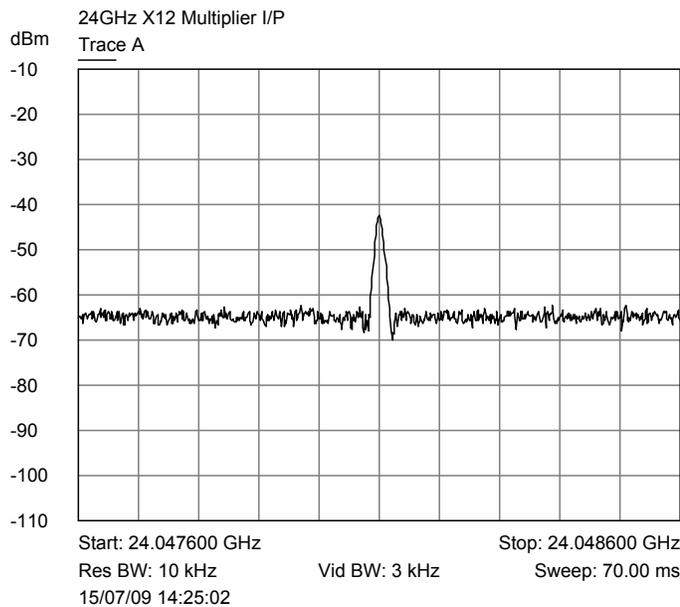
Dynamic range -20 to +13dBm. Measurement was simply made by connecting a DVM onto the detector SMA output. Input via an SMA wg-42 transition.

On 47Ghz output was 0.15V for around -10dBm in.

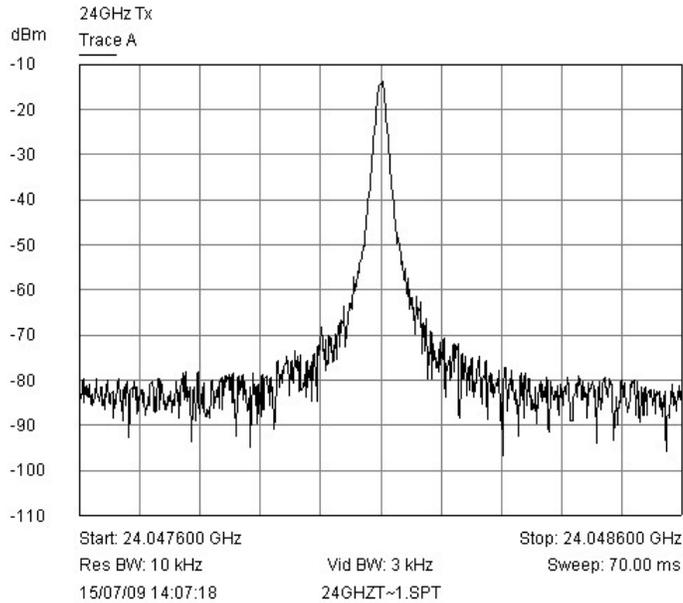
Use as a Multiplier

The multiplier input can be driven up to about 2.1GHz, drive +10 to +17dBm. A multiplication of X12 worked well for 24GHz. The drive level required for optimum output changes with frequency / multiplication. It is not linear with power, expect peaks around +10, +14, +17dBm input.

The detector input, can be driven as a multiplier up to at least 5Ghz level +8 to +17dBm. Output is good at 24GHz, and detectable up to 5 metres away on 47GHz (I used a multiplication of X23), and should be detectable at 76GHz. In practice it is easy to try each of the 2 inputs, and see which is better for the output required. Although at 47GHz the 'det' input definitely gave the best results.



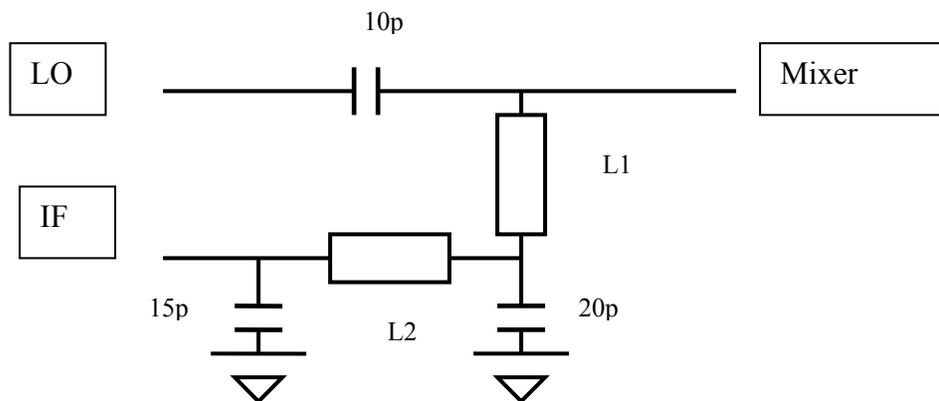
Use as a Harmonic Mixer



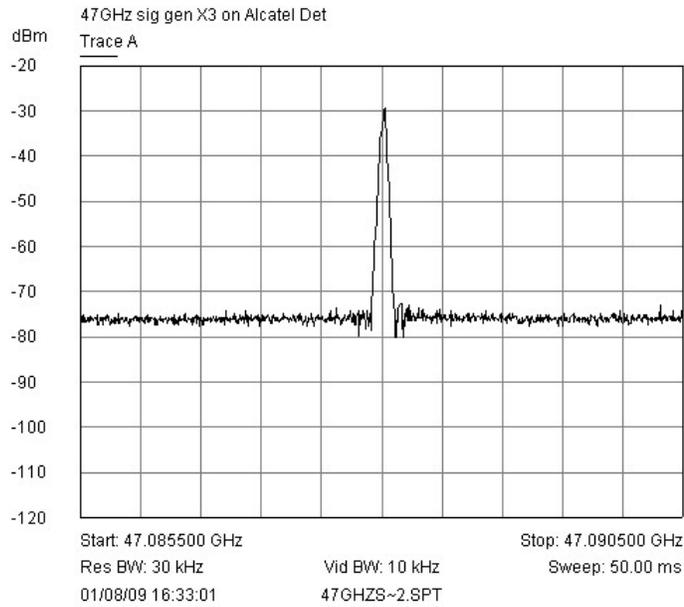
The plot shows the module being used as a harmonic mixer on an Advantest R3271 spectrum analyser. Although the analyser will work at 24GHz, in this case it is in external mixer mode. A single connection is used between the analyser and 'Det' SMA on the module. LO drive and IF input are on one coax, as the analyser contains an internal duplexer. Trying to use LO on the 'mlt' input, and using 'det' as an IF output did not work.

For use with an HP8566A or later analyser, a duplexer will be required. A simple duplexer is shown below. I made the IF filtering to cut off around 450MHz. The construction is more important than the actual component values. The inductors were about 8 turns 3mm dia. of fine wire, pulled around to get the best performance at 432MHz. I added a miniature ferrite on L1 to damp a resonance.

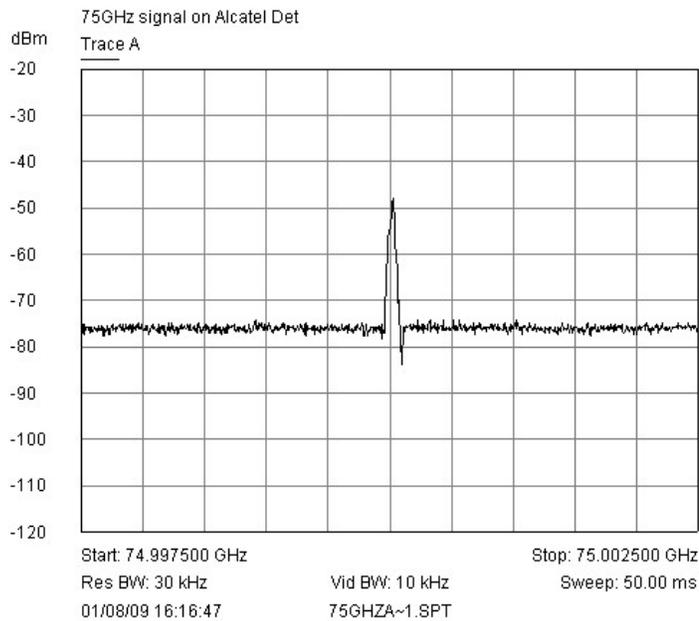
Block diagram



LO / IF Duplexer for HP Analysers



Harmonic Mixer with 0dBm 47Ghz input



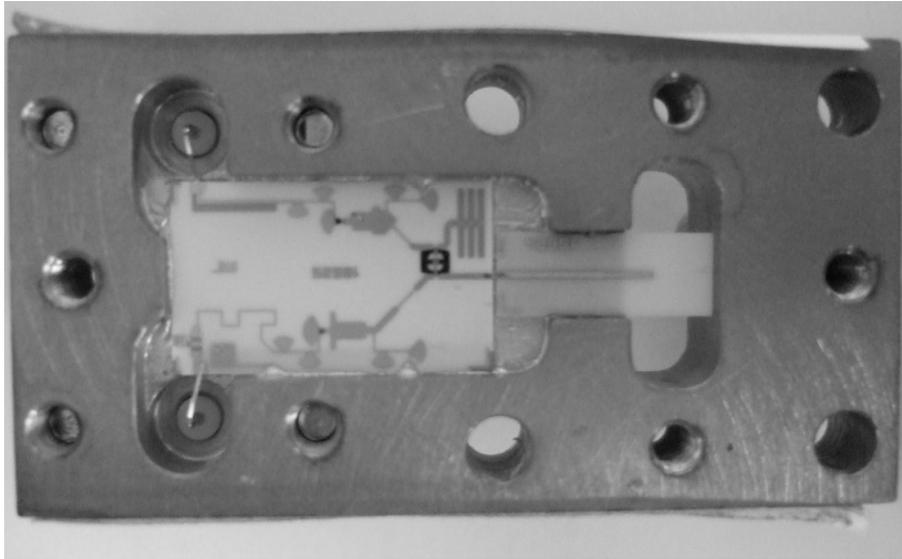
Harmonic Mixer with 75GHz input

I was surprised how well it worked at 75GHz (it works at 76GHz but my X5 source drops off 10dB unfortunately). Input was via 4 inches of wr-10 waveguide, just to make sure it really was 75GHz. It would be fairly easy to add a tuning screw in the back panel to see if better performance at specific frequencies could be obtained. Using it as a mixer and external duplexer I could listen to my 47Ghz transmission using 432MHz as an IF.

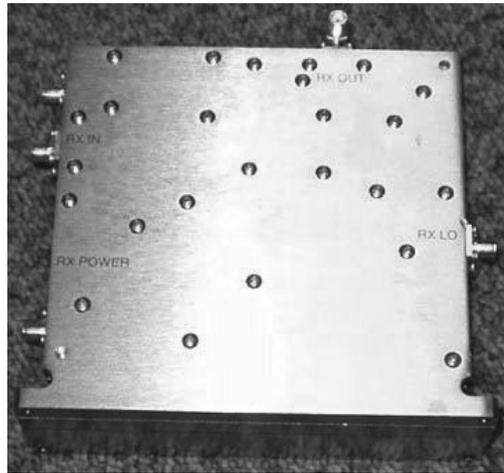
The plots taken on the R3271 analyser are with approx. 15dB internal attenuation removed, as it was converted to remove the internal diplexor.

Internal construction

– the small black blobs are the diodes, the bit looking like 2 back to back Ds in a dark background is the coupler. The printed inductor – bottom left, forms a LPF limiting the input frequency of the ‘MLT’ input.



www.g4jnt.com/whiteboxmods.pdf



Ceragon 7GHz Module for 5.76GHz by Roger Ray G8CUB

These will work after modification on 5760MHz. They consist of a receive LNA mixer & IF amplifier, plus a transmit filter, amplifier & output monitor. Thus just needing an LO, Tx mixer, and a couple of relays to complete a 5.76GHz transverter.

Measured performance on receive with a 432MHz IF, is as follows:

Conversion gain +19dB

System noise figure 2.6dB

Image rejection (5.328GHz LO) -22dB

Performance on transmit:

+34.4dBm output (2.75W) - 1dB compression

+35.2dBm output (3.3W) saturated

+57dB gain

Supply requirements:

	Receive (Tx Inhibited)	Full Output
+5.0V	0mA	100mA
+6.8V	180mA	0mA (switched)
+6.0V driver	0mA	0.93A
+10.0V pa	0mA	0.95A
-6.0V	40mA	40mA

On the transmit side, the only essential modification is to place a piece of ceramic on top of the existing inter-digital filter. This shifted the entire frequency response down about 1.5GHz, to exactly where needed. Transmit image was 60 – 80dB dependant on the size of the ceramic. With the high transmit gain, it is necessary to either use an input attenuator, or use the AGC control to turn the gain down. This will give 1dB compression of around +32dBm. Adding some small tabs around the PA pushed the power up to +34.4dBm.

On receive, the image filter strips were extended by about 2mm using silver paint. This gave 22dB image rejection, but conversion gain was only about 11dB. Starring the RF amplifiers with small pieces of copper, improved the gain by about 7dB. The IF response was about 1 dB down at 432MHz, so the IF output capacitor was changed to 330pF to give a final conversion gain of 19dB. On a second unit, I used a piece of ceramic like on the TX filter. However the size is critical, it needed to be very slightly less than the board outline of the filter. With a bit of playing, I achieved 20dB conversion gain (modifying the RF amp as above) with the image -25dB. Also on this one, I was not convinced that the IF capacitor change made any difference, so it may be easier to leave it alone.

In my 5.7GHz system I use the Ceragon block with an Alcatel synthesiser switched for either 10 or 5.7GHz, using a 10MHz reference.

No information was available on pin-out or supply voltages, so the following is my derivation. From data on the TX driver, a negative supply of -6V is required.

Pin-out on the two connectors as pictured above, left to right, is as follows.

Rx

Connector 1 (6 way)

1	2	3	4	5	6
n/c	+6.8V	-6V	0V	+6.8V	n/c
	IF			RF	

Tx

Connector 2 (6 way)

1	2	3	4	5	6
n/c	+6V	-5V	0V	+5V	+10V
	driver			amps	PA

Connector 3 (5 way)

7	8	9	10	11
n/c	Tx Det	AGC	0V	Tx Mute

Connections to the unit are by two, 0.1” pitch single in line connectors. It is quite easy to solder wires straight on if you don’t have suitable mating plugs.

In my system, the 6.8V supply is provided by a 7808 regulator and dropper resistor. The 10V supply, uses a 10V 1A low voltage drop regulator. The -6V is provided by a block DC-DC converter, and -6V (79L06) regulator.

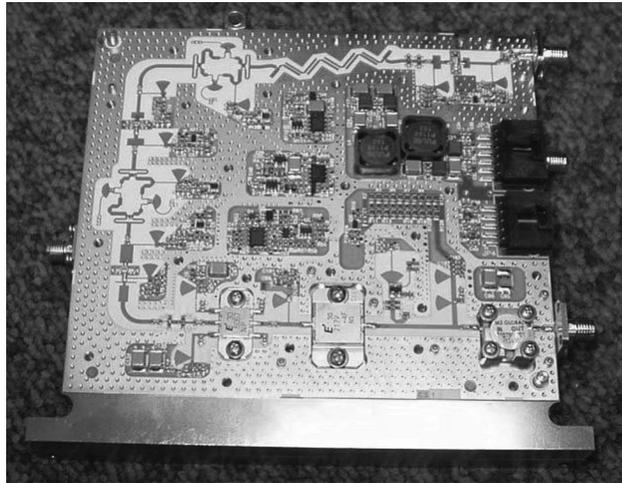
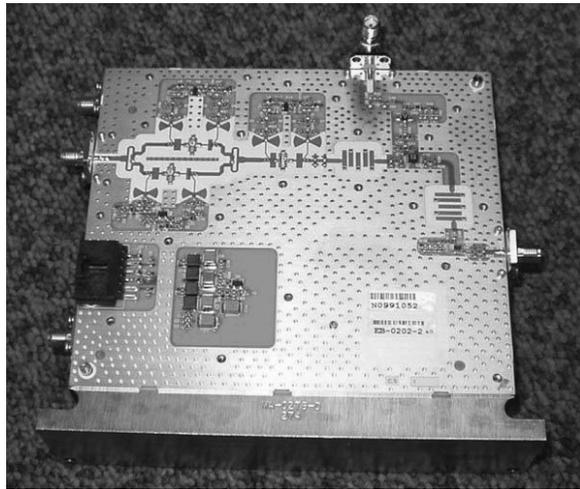
External protection is required, to inhibit the positive supplies, if the -6V supply fails.

Tx Mute – 0V to inhibit

Tx Det – DC proportional to dB output power (log detector) Max ~ 4V

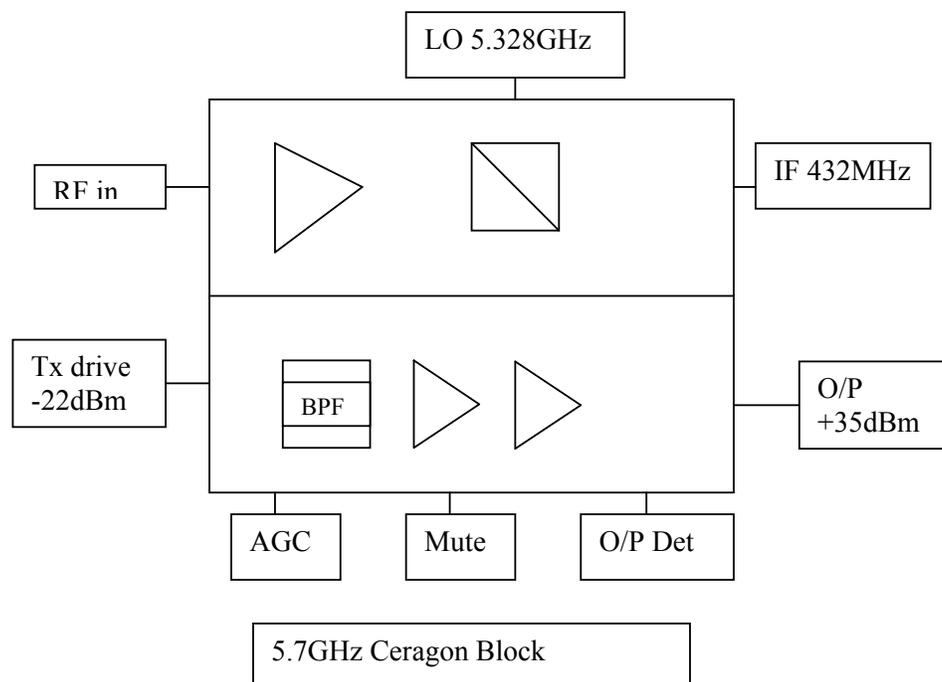
AGC – 0- 5V Control I guess (I have not tried this, just left the pin o/c)

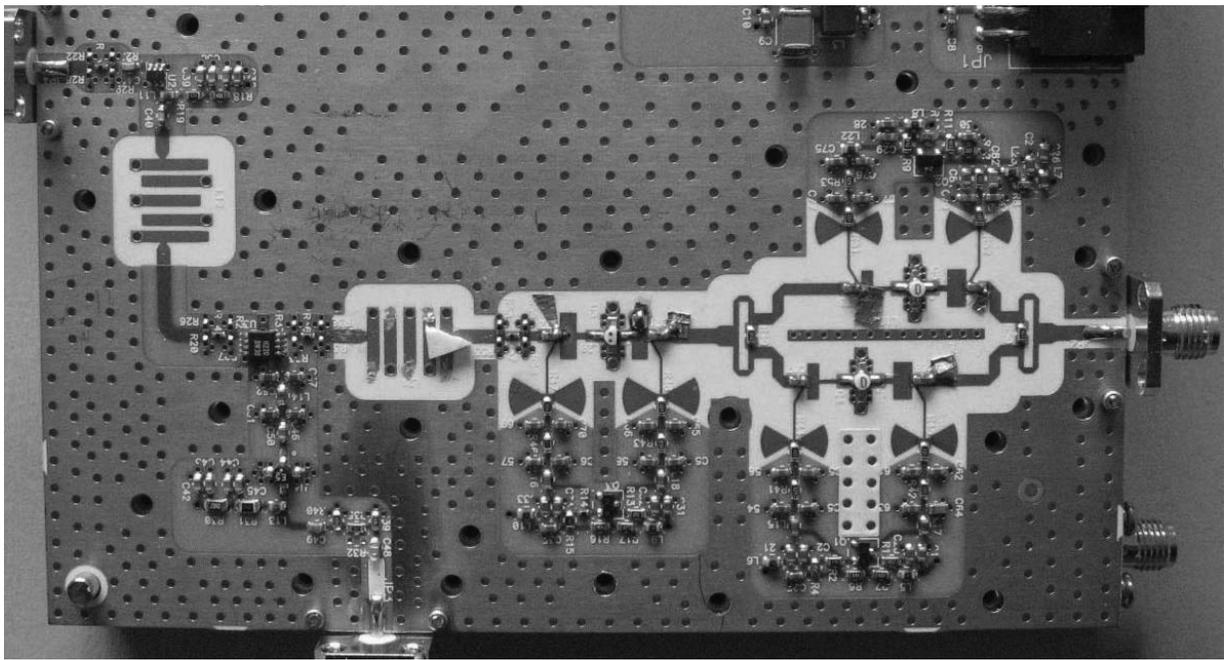
Local oscillator power required is +10dBm (5.328GHz). Around -22dBm Tx drive will give you full output (at max gain). The 2 SMA output connectors allow easy use of a coax relay, or the addition of a single stage low noise amp and / or PA.



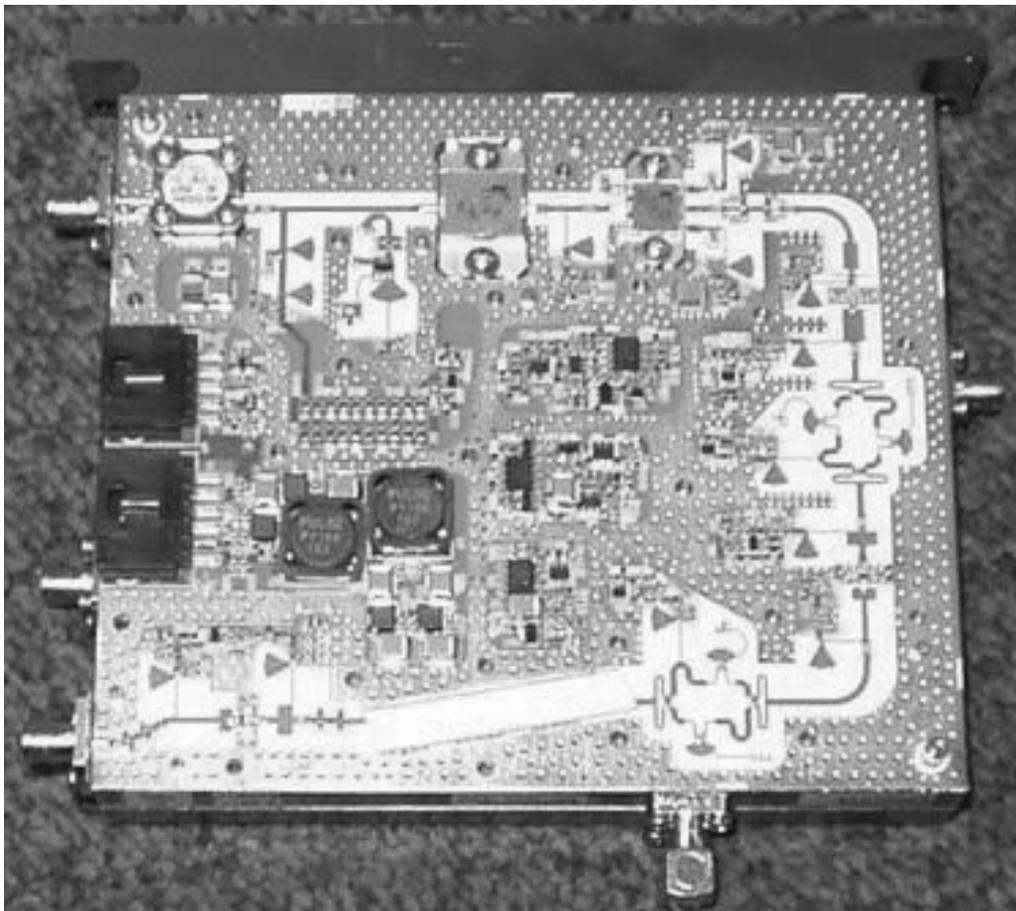
Unmodified PCBs

Block diagram

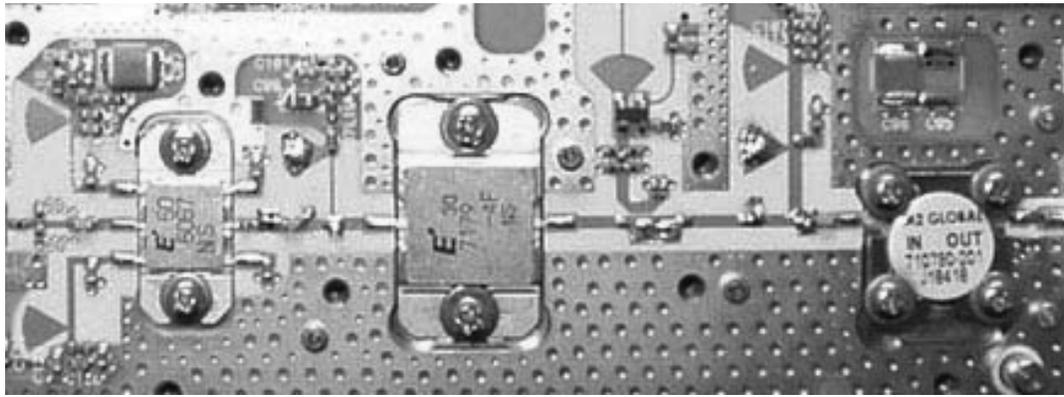




Modified receive board



Modified Transmit board – ceramic in place.



Power Amplifier mods for 2.7W Output. Driver FMM5057VF, PA FLM7179-4F,

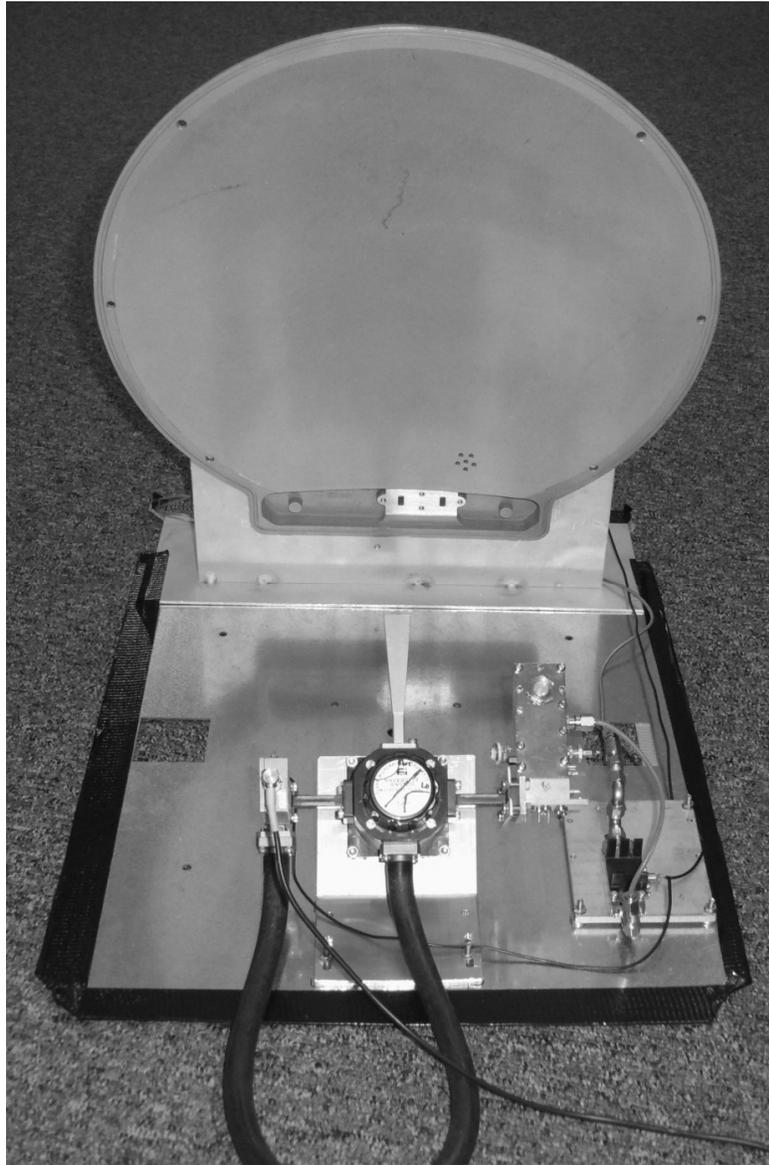
Availability

Units are currently available on Ebay from ‘art-in-part’. These however may have some faults, particularly damaged SMA connectors (an easy change).

I have a few that have been checked, but are un-modified. These are available on a first come first served basis for £40 plus £4.50 postage (UK). These will be provided with Tx ceramic strips. Contact littlemallards@hotmail.com

Simple 47GHz Transverter

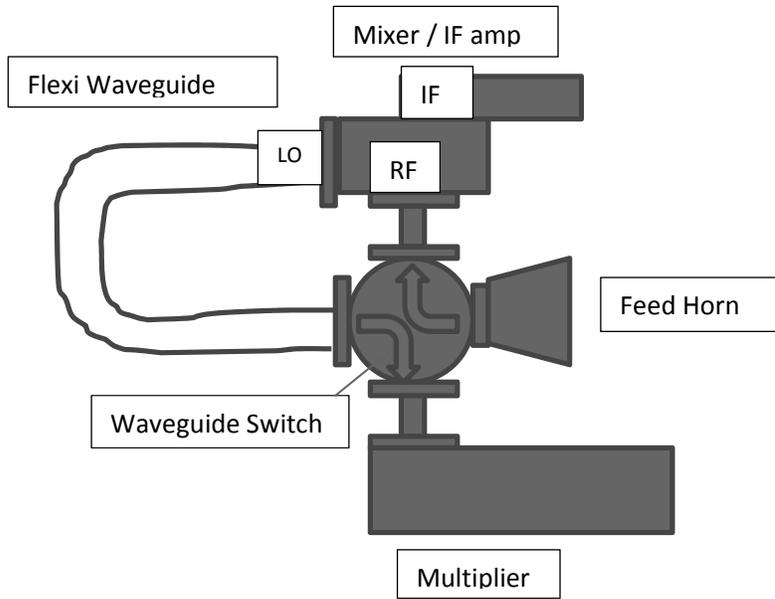
By Roger Ray G8CUB



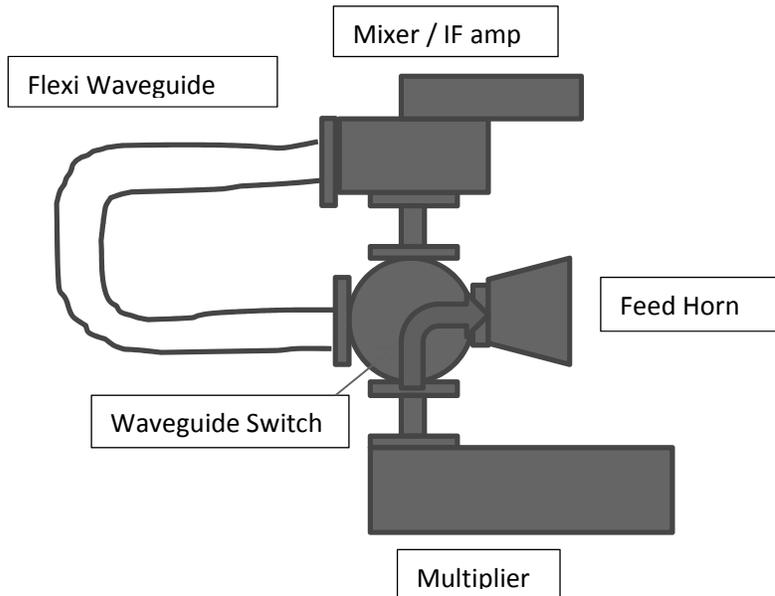
The idea of putting together a quick transverter for 47GHz came to me after 2012 RAL. Having come away with an offset dish (thanks Peter), and a WR-28 four port waveguide switch. It was decided to try and put something together in 4 hours, and have it ready for the '76th' August tests. Of course it took 6 – 7 hours to build, but was finished in time!

The other component that I needed had come from ebay - A Honeywell WR-28 mixer, with IF amplifier attached. I had previously removed the IF amp, and was using it as a X2 multiplier 38 / 76GHz. It worked pretty well in that role giving 1.2mW out for 80mW drive. Hopefully it would work well in its original role as a mixer, but at 47GHz.

From my original Pasolink transverter, I was retaining just the X4 multiplier. This was to be driven at 11.7 GHz from an Elcom synthesiser via an amplifier to achieve sufficient drive. The same synthesiser to be used as LO for the mixer, and TX output.



Waveguide Switch in Receive position.



Waveguide Switch in Transmit position.

The constraints of the 3.33MHz steps with the Elcom, meant using an odd IF of 434.66 – 436.66 MHz, as I wanted to keep the LO on the low side of the RF. Although that was not a problem using the FT817 as IF receiver. The biggest problem was getting a TX signal between 47,088 and 47,089MHz. The nearest synthesiser output was several MHz away. My solution was to use an offset synthesiser reference of 9.999 MHz for TX, while keeping 10MHz for RX. The synthesiser was programmed for 11.663333 for receive, and 11.773333 for transmit. The transmit frequency is then $(11.773333 \times 4) \times 9.999/10 = 47.088624$ GHz.

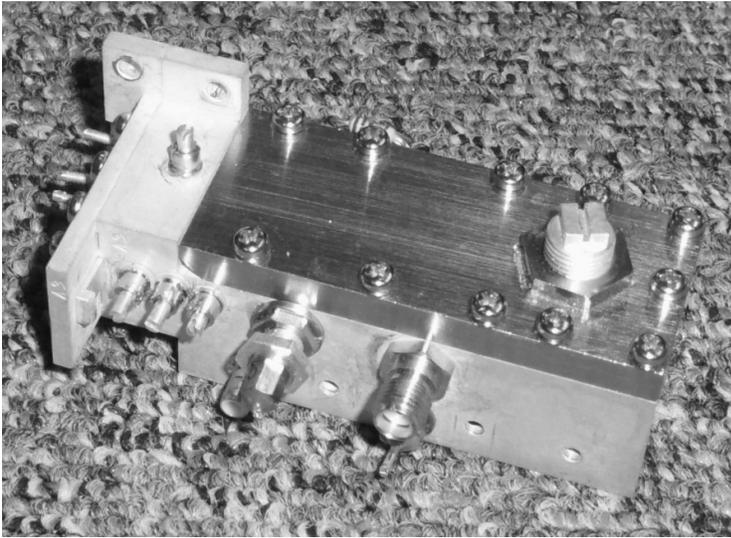
An old crystal oscillator with a heater was to hand, being part of a paging transmitter. The Ovenaire heater was for 85 deg.C I had a Quartz Lab crystal for 9.9921MHz 70 deg.C, but it was of the right order, and near enough in frequency to be pulled.

The oscillator had a varicap which might allow FM, if the FM would pass the pass through the synthesiser OK. A quick previous trial with a TCXO was unsuccessful, as the frequency shift was limited, and it refused to FM, probably due to internal decoupling.

The first stage of the build process was to program the PIC for the two frequencies, and modify the Elcom for external reference. The synthesiser that I had, came with reference in / out connections on the side, ideal I thought. However I could find no way of making it change to external reference input. I concluded that it must be a software command. In the end I just used one of the connections to give a wired reference input connection. Any attempt to keep the internal 10MHz reference running produced a signal like a Christmas tree, when external ref. was used. So in the end I reluctantly used an external 10MHz reference and the 9.999MHz offset ref. switched with a pcb coax relay. To drive the Pasolink multiplier I needed 100mW or so. To get this level I attenuated down the Elcom output and used an Avantek amplifier that would give the power.

Dave G4FRE has an article on his web page about modifying the Pasolink
<http://www.g4fre.com/pasolink2.pdf>

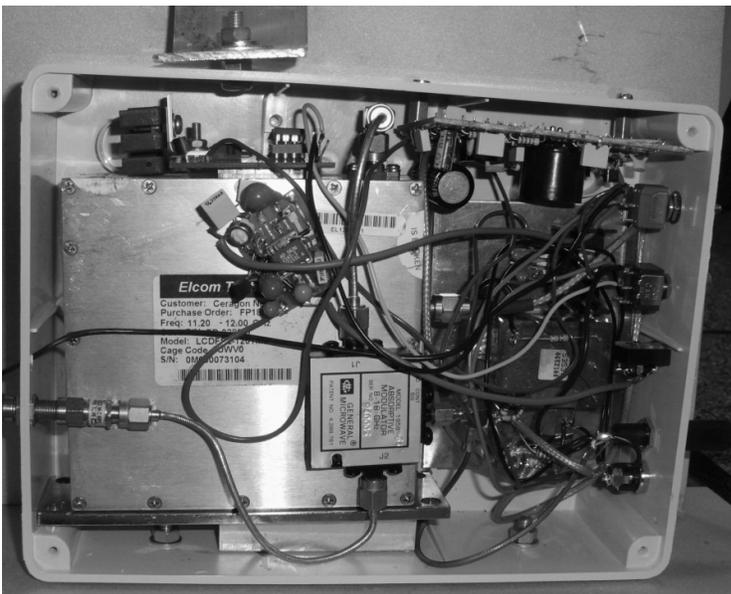
In this case I was just using the x4 multiplier, so alignment was straightforward compared to the multitude of small tuning screws on the whole system! To connect to the output I butted up a WR-28 waveguide flange, and made up a couple of metal strips to hold it in place. Modification is, remove the SMA monitor connector, and remove the spacer below it and screw it back in. Once this is soldered, and the original resonator removed, the SMA becomes the 11GHz input.



Pasolink multiplier from 50GHz unit.

Tuning is by the 4 large screws, and the 3 small ones. However most significant is the slider, under the over-hanging top. This can easily be missed. The screw above it locks it in place. The bias pot is set for maximum output, I measured 5.6V across the it.

I have always struggled to get the claimed output from these multipliers. In this case by upping the drive to +22dBm, I achieved an output of +6dBm (4mW). There is a good rejection of the 23 & 35GHz harmonics, but it may be best not to look too closely at the 58GHz component.



To allow for CW keying, I used an inline 'modulator' which switches the synth. output. FM was very last minute. I wanted some audio amplification and limiting. The only thing I found was a compressor chip, that I had never got around to using. This was stuck on the synth. and bias provided for an electret microphone.

The synthesiser and references were crammed into a plastic box – not pretty but it works. The dish, in-line amplifier and waveguide switch were mounted on a base plate as can be seen in the pictures.

The feed horn was hack sawed through to give the right spacing to the dish. It is likely this under illuminates the dish, but time was running out.

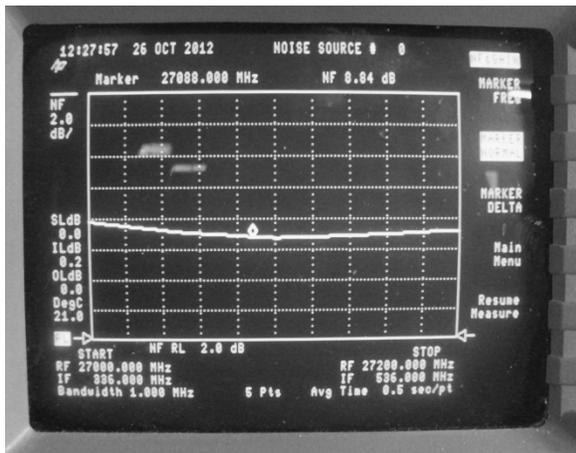
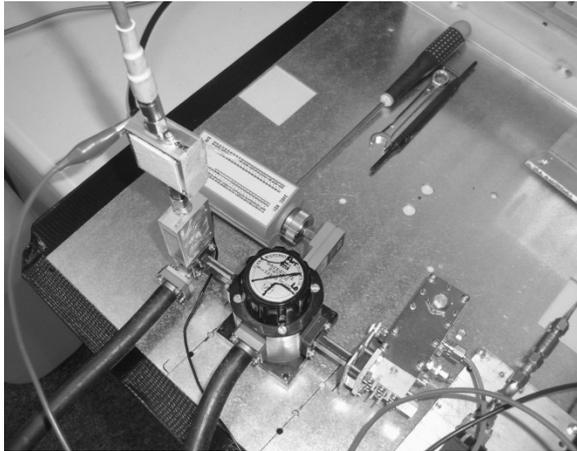


I had previously only had one QSO on the band, with Harold G3UYM/P using a Pasolink system, at the great distance of 100 metres.

The first contact with the transverter was at Bignor Hill, both Peter G3PYB and Ian G8KQW were 59+ at 53km on the Isle of White, with similar reports in the other direction using FM. It was great to know all was working. Dish alignment was not that critical, which probably meant that the feed position was not optimised. Also Ian's comment on the CW when in beacon mode, suggested frequency pulling when in CW mode. Not surprising in that the oscillator had a basic zener 10V regulator, and the keying was causing a significant shift in amplifier current.

58 / 59 FM reports at Ditchling Beacon @ 83km showed the transverter to be working well.

Later noise figure measurement, showed that the WR-28 flexi-waveguide was poor or damaged. When replaced a noise figure of 8.8dB was measured at the input to the waveguide switch, showing the mixer to be working pretty well at 47GHz.



Noise figure measurement with a HP346C K01 source. For 47GHz read 27GHz as the 8593E analyser allows a max. frequency offset of 30GHz!

ICOM IC-706Mk2G injection locking

Injection locking of the 30 MHz
reference oscillator using a 10 MHz
oven controlled oscillator.

SM6FHZ 2018-04-29

Prerequisite and short description

- ICOM IC-706Mk2G has not a reputation of being very frequency stable. This limits the use of it for certain modulation schemes and modes. The drift is a problem mostly at the VHF and UHF-bands of the IC-706Mk2G, especially on 432 MHz where it can be quite substantial over the Tx-Rx-cycles when the fan goes on and off.
- The ICOM IC-706Mk2G uses a 30 MHz oscillator to derive all LO-frequencies in the rig. This opens up for a quite easy way of locking all LO's in the rig.
- A few references were found on the Internet (see references below) on the possibility to injection lock this oscillator with a high accuracy external signal. I experimented with a few possibilities to do this and ended up in the below presented solution.
- I found that the square wave output of a 10 MHz oven controlled oscillator contained quite a high level of 30 MHz signal. I decided to filter out this 30 MHz signal to improve the spurious rejection and enhance the 30 MHz signal purity.
- An SMA connector mounted on the rear side of the IC-706Mk2G and a RG-316 cable routed to the 30 MHz oscillator shielded box inside the rig. A small coil couples the injection signal to the tank circuit of the internal 30 MHz oscillator. Putting a 10 dB attenuator on the external reference signal input improves isolation and ensures that the IC-706Mk2G can be used without an injection signal if needed (then with the original frequency stability).
- Voltage for powering the 30 MHz external reference unit is taken from the Acc-connector of the IC-706Mk2G.
- Warm-up time for an acceptable frequency accuracy and drift is about 2 minutes in my case.
- This frequency locking scheme has worked well for me when running 432 MHz meteor scatter skeds as well as copying digital mode signals via the Moon.

About Injection locking

- Injection locking of oscillators is not an new idea or technique. It has been used for many years.
- One common usage has been to lock e.g. magnetrons, Gunn and IMPATT oscillators at microwave and mm-wave frequencies. Here a circulator is commonly used to inject the control signal. I designed injection locked IMPATT oscillators at Ku-band (at QRL) in the early 80's.
- You can in most cases expect a gain of about 10 to 15 dB for a reasonable locking band width. In the case described in this presentation the requirement of the locking band width is quite small, it only has to accommodate for the temperature drift of the master oscillator of the IC-706Mk2G in the environment you are in. The phase noise of the locked oscillator is set by the control signal over the locking band width, so a good locking band width is beneficial as long as the reference signal is better than the unlocked oscillator.
- More on injection locking can be found here
 - <http://www.seas.ucla.edu/brweb/papers/Journals/RSep04.pdf>
 - <https://ntrs.nasa.gov/archive/nasa/casi.ntrs.nasa.gov/19880003345.pdf>
 - https://en.wikipedia.org/wiki/Injection_locking
 - http://rfic.eecs.berkeley.edu/~niknejad/ee242/pdf/eecs242_lect26_injectionlocking.pdf

Use the ready made hole on the back side of the IC-706Mk2G, that is covered by a washer, to mount a SMA-connector



Route the double shielded Teflon coaxial cable, RG-316, to the 30 MHz oscillator box. Enter the box through the ready made hole that fits the inner dielectric of the RG-316.



Solder the shield to the outside of the oscillator box.



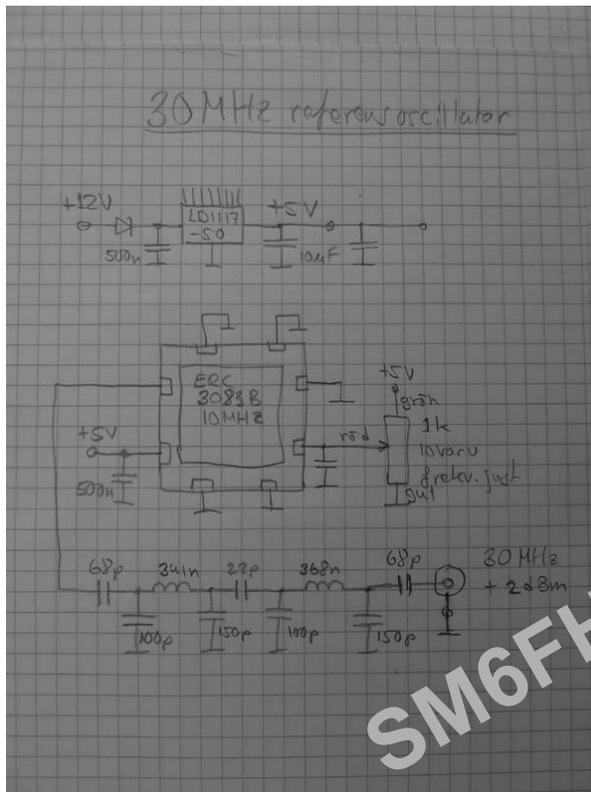
Wind a 6 turn coil that fits on the L601 coil in the oscillator box.



Put a 10 dB SMA-attenuator on the input connector for isolation towards the outside world

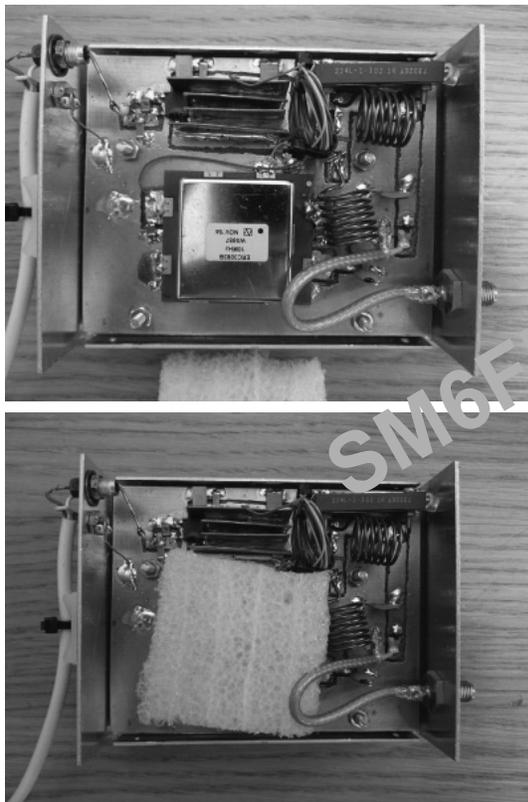


Schematic of the 30 MHz reference oscillator



- 10 MHz oven controlled X-tal oscillator ERC 3083B (that I used). Can be found on Ebay. Other oscillators may work as well.
- +12V supplied from ICOM IC-706Mk2G and stabilized to +5V.
- Frequency adjustment is facilitated by the use of a 1 kohm 10-turn potentiometer.
- A band pass filter at 30 MHz filters out the 30 MHz frequency component from the oscillators square wave output. Measured level at 30 MHz on my oscillator is +2 dBm.
- Specified warm-up time for the ERC 3083B is <6 minutes at room temperature to reach +/-0.1 ppm from final frequency. My oscillator needs ~2 minutes to reach a acceptable frequency accuracy and stability on 432 MHz.

Interior of reference oscillator



- The oscillator is housed in a small metal box with filtered +12V voltage input to the upper left. SMA 30 MHz output connector in the lower right corner and frequency adjustment potentiometer in the upper right corner.
- The +5V regulator is cooled using four small copper sheet metal pieces soldered to the large pad on the PCB where the regulator output is connected.
- The 30 MHz BPF can be seen to the right of the oscillator.
- The oscillator has a small foamed plastic thermal insulator hat to enhance frequency stability and slightly reduce oven supply current (lower picture).
- The PCB pattern is made from a double copper sided FR4 board by using a Dremel with a 1.6mm engraving bit. Surprisingly simple and fast process. Not a very beautiful art object, but fully functional.

Performance

- The 30 MHz oscillator in my ICOM IC-706Mk2G locks to the injected signal from a level of -20 to -15 dBm. That means -10 to -5 dBm is needed from the external reference oscillator when the 10 dB attenuator is used. A small margin in input signal is good to have to ensure good locking at all times.
- Using a 10 MHz TTL compatible square wave without anything else I get a good lock using the 10 dB attenuator. Checking it using a spectrum analyzer showed that the 30 MHz component was about -12 dBm (on my 10 MHz working reference used for my frequency counter and signal generator). This may vary a lot from oscillator to oscillator.
- It would be nice to use a 10 MHz to 30 MHz filtered tripler as show in the references. However, experiments showed that the conversion loss of the tripler (10 MHz to 30 MHz) was so high that the 30 MHz component of the square wave directly from the oscillator gave a better signal at 30 MHz than using the tripler. The tripler contains filters and does not benefit from the oscillators signal level at 30 MHz for it's output. I did end up not using the tripler after all. If you use a 10 MHz reference oscillator with a sine wave output the tripler solution might be the way to go.
- The frequency locking scheme used proved to be a very simple yet efficient way of locking the ICOM IC-706Mk2G to a external 10 MHz reference.

References

- http://www.vk3hz.net/XRef/XRef_Icom_IC-706MKIIG.pdf
- <http://www.wenzel.com/documents/2diomult.html>

SM6FHZ

Dish Extension - 3m to 4m Parabola-Stress Hybrid Construction Techniques

Paul Andrews, W2HRO
March 2019



North America & 3m Dishes

- North American is the birth place the 3m TVRO
- 2.6 Million 3m dishes were installed in the USA.
- Abandoned 3m dishes are usually available for free
- EME Amateur should move fast before they are gone



Lots of 3m Dishes - but few larger

- The vast majority of US C-band dishes are 3m (10 ft.)
- 3.7m (12 ft.) C-band dishes were used but rarely.
- I wanted to extend my 3m dish to 4m.
- The extension needed to be cheap, light and effective.
- I'll use this dish on 432 and 1296 MHz bands.



3m dish

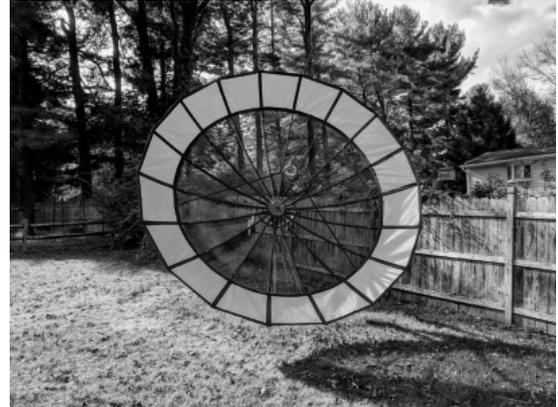
Compare 3m vs 4m Dish

- The gain of a dish is related to its size or effective aperture area. A bigger dish has more gain.
- Extending from 3m to 4m nearly doubles performance.
- 3m dish = 4.2 sq. m and 4m dish = 7.5 sq. m
- At 432 MHz / 3m = 20.5 dB gain / 4m = 23 dB gain
- At 1296 MHz / 3m = 30 dB gain / 4m = 32.5 dB gain
- A bigger dish has a more narrow beam width and reduced noise pickup.

3m Dish extended to 4m



3m Dish



4m Dish (extended)

Total Project Cost - \$100 (approx.)

Parabola – Stress Dish Hybrid

- My extension is not a true parabola
- My extension follow the curve of a “stressed” rib.
- K2UYH Stress Dish Design - [http://www.nitehawk.com/432 MHz EME/stress dish.pdf](http://www.nitehawk.com/432_MHz_EME/stress_dish.pdf)
- The 3m to 4m extended dish is a hybrid design.
- True parabola for the first 3 meters
- Stress dish from 3m to 4m
- A stressed rib deflection at the edge of a dish is very close to a true parabola.

Stress Extension Design

- The circumference of a 3m dish is 30.9 ft.
- The circumference of a 4m dish is 41.2 ft.
- My rib extension were sized to support a new 41.2 ft. circumference outer edge of the dish
- The 3m dish had an $f/d = .37$
- The focus point of the dish does not change.
- The 4m dish has an $f/d = .28$
- The bigger dish is now much deeper.
- The deeper dish is nicely illuminated by a Patch Feed.

Extension Rib Materials

- Kevlar String
- Notched End-caps
- 6mm 36" arrow shaft
- Battery terminal lug
- 5mm aluminum rod
- Aluminum-clad Mylar



99 % Aluminum-clad Mylar



NASA ECHO-1
Passive-reflector
Satellite



Radiant
Attic
Insulation

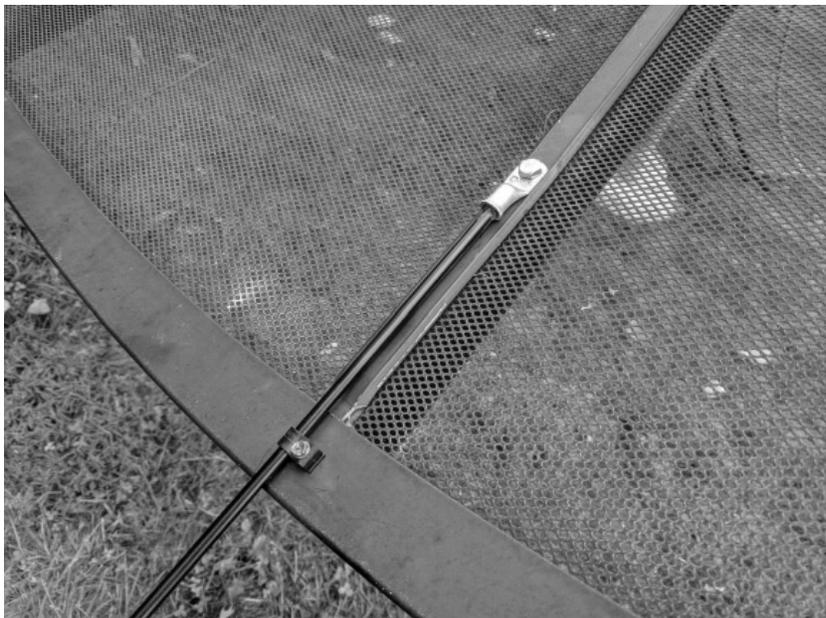


99% Aluminum-clad
Mylar (rip-stop)
White Color (one-side)

Extension Rib Installation

- 36" arrow shaft
- Glue on terminal lug
- Cable clamp
- Battery terminal lug

Each dish designs may require a different rib attachment technique.



Outer Perimeter String Installation

Attach the 40.2 ft. of Kevlar Perimeter String to end caps

Maintaining an accurate circumference is required

Use UV stable duct tape to attach Qty. 18 sections of 5mm aluminum rod to stiffen the outer edge.



3m dish with 4m framework – no Mylar covering.

Aluminum-clad Mylar Panel Installation

Mylar is cut with scissors into panels – Qty. 18.

Each panel is attached to the stress frame with tape

Gorilla Glue-brand tape is waterproof, UV-stable and really sticky and strong

I installed all 18 panels in about six hours.



3m dish with 4m framework – partial Mylar covering.

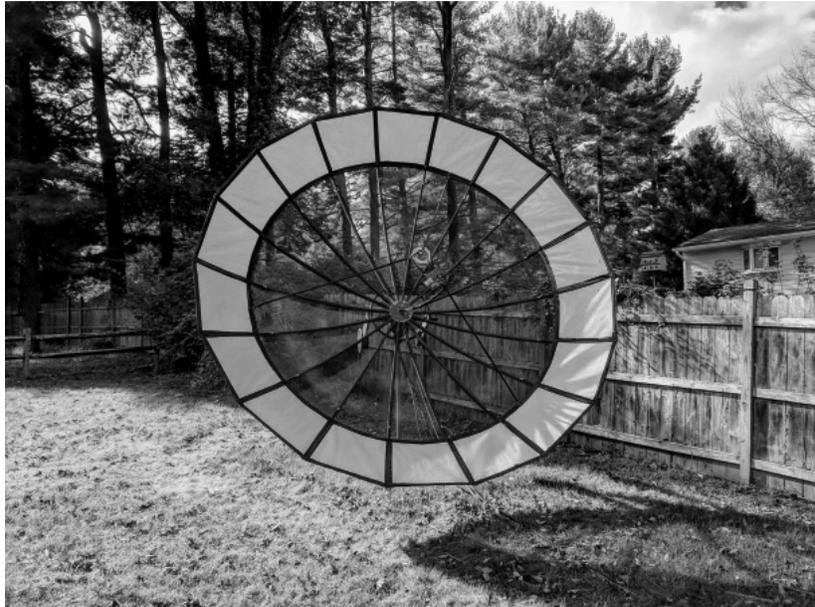
Completed 4m Dish (extended)

Total construction time
16 hours over several days
and I remained QRV.

Weight Increase – 6 lbs.

Wind load - Yes. The solid
panels do increase wind-
load. But the arrow shafts
flex and shed the wind.

The dish survived the FN20
winter of 2019 with no
problems.



Completed 4m dish with completed stress extension.

4m Hybrid Dish Performance

- 1296 MHz sun noise increased from 10 to 12 dB
- 432 MHz sun noise increased from 5 to 7 dB
- 1296 MHz JT65 signal reports are very good
- Completed 432 MHz WAC with loop feed and 180w.
- Can worked a 1296 MHz single Yagi with 120w.
- Dish pointing on 1296 MHz is more critical.
- 432 MHz receive performance is exceptional.
- Removed BFR from 1296 MHz Patch for better illumination
- Removed cake pan baffle from 432 MHz loop feed

Credits To my Elmers

- K2UYH – Stress Dish Design

[http://www.nitehawk.com/432 MHz EME/stress dish.pdf](http://www.nitehawk.com/432_MHz_EME/stress_dish.pdf)

- SM6FHZ / SM6PGP – 23cm Patch Feed

<http://moonbouncers.org/>

[Circularly%20polarized%20patch%20feed%20for%201296%20MHz
A.pdf](http://moonbouncers.org/Circularly%20polarized%20patch%20feed%20for%201296%20MHz_A.pdf)

- OK1DFC – 70cm Loop Feed

<http://www.ok1dfc.com/eme/technic/432feed/432feed.htm>

Summary & Next Steps

- My initial goal was to improve 432 MHz performance, but I see good benefits on 432 & 1296 MHz
- The biggest single improvement was 432 MHz receive
- Measurable sun noise increase on both 432 & 1296 MHz
- Fast, Cheap and Effective dish upgrade
- The upgrade can be completely reversed in a couple hours

- Try to build a portable dish using Aluminum-clad Mylar
- Try 144 MHz EME with a 4m dish

23cm EME - Patch Feed Construction and Mounting Techniques

Paul Andrews, W2HRO
September 2018



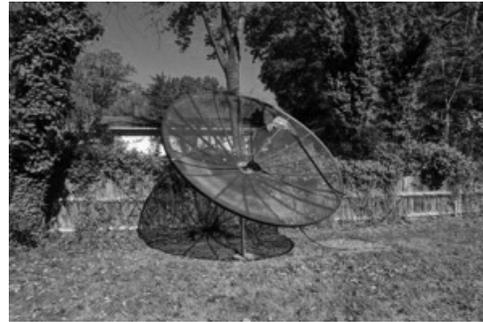
North America & 3m Dishes

- North American is the birth place the 3m TVRO
- 2.6 Million 3m dishes were installed in the USA.
- Only 150,000 C-band dished are still in use.
- Abandoned 3m dishes are usually available for free
- EME Amateur should move fast before they are gone



My Path to 23cm QRV

- My neighbor offered his 3m dish if I removed it
- The dish was 30 yrs. old but in excellent condition
- I devised a simple way to mount it to my SPID RAZ
- What feed to use and how to mount in on a TVRO?



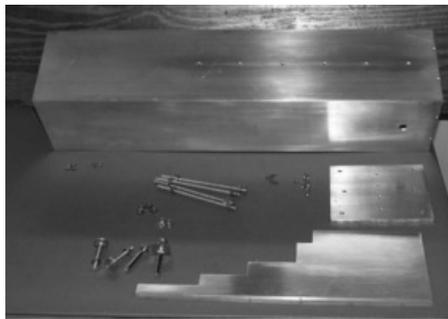
Easy Decisions – Station Setup

- I have a TS-2000X with the 1296 MHz module
- VHF Designs 150 SSPA / WA2ODO 0.3 dB NF preamp
- Feed line / sequencer / isolation relays / WSJT-X
- SPID RAZ – AZ / EL Rotor with 1 deg. resolution
- Feed? Septum or Patch

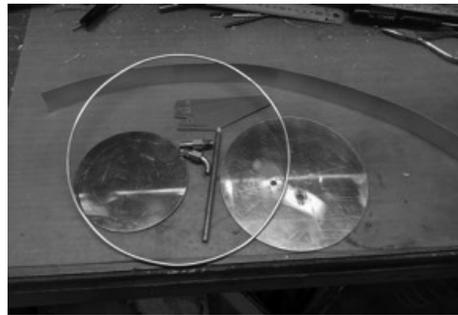
Difficult Decision (for me)

- I have a KL6M 23cm band Septum (unassembled)
- The Septum feed is the gold standard
- The TVRO feed support will not hold a Septum
- I learned about the SM6FHZ / SM6PGP Patch feed

KL6M Septum



SM6PGP CP Patch



Septum vs. Patch Feed

- The Septum feed has superior performance potential
- The Septum feed is bigger and heavier
- The Septum feed can handle a 1kw of power
- The Patch feed has good performance potential
- The Patch feed is smaller and lighter
- The Patch feed is native dual pol linear
- The Patch requires an external hybrid divider for CP
- My Patch feed is limited to 250w max power

Septum Feed Mounting

- The Septum feed is bigger and relatively heavy
- The Septum feed requires robust support legs
- The typical TVRO feed support is not compatible

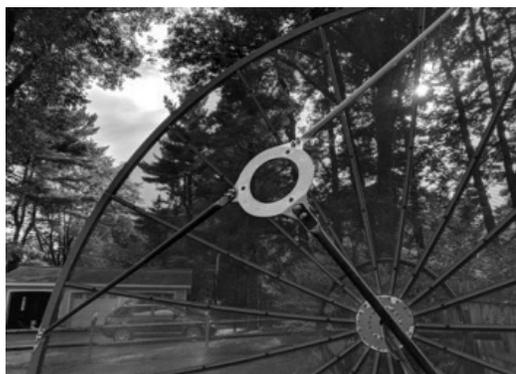
RF Ham Design Feed Mount



Typical TVRO Feed Mounting

- The TVRO Dish is design to support a Chaparral feed
- The Chaparral feed has scalar ring mounting points
- Repurposing the original TVRO feed support is EASY

Typical TVRO Feed Mounting



The “Eureka” Moment

- The Chaparral and Patch feed are the same diameter
- A modified Chaparral will hold the Patch feed elements
- This Patch feed design mounts directly to the TVRO
- The focus point is correct for the Patch feed

Chaparral Feed in TVRO



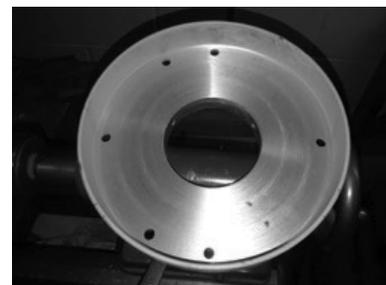
Patch Feed in TVRO



Modify the “Chaparral” Feed

- Find a Chaparral C-Band choke (aluminum)
- Removed the inner rings with a lathe
- The inner diameter of the Chaparral is 160mm

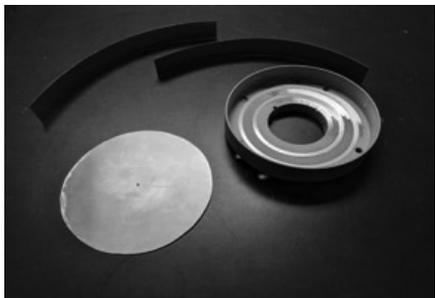
Remove the Inner Rings in Chaparral C-Band Feed with Lathe



Building the “Chaparral” Patch

- Chaparral choke become the foundation of the Patch
- The inner diameter of the Chaparral is 160mm
- The Patch reflector is 160mm and fits inside the choke
- The Patch baffle is glued to the inside of the outer ring

Choke, Patch Reflector & Baffle



160mm Diameter



Baffle glued to Choke



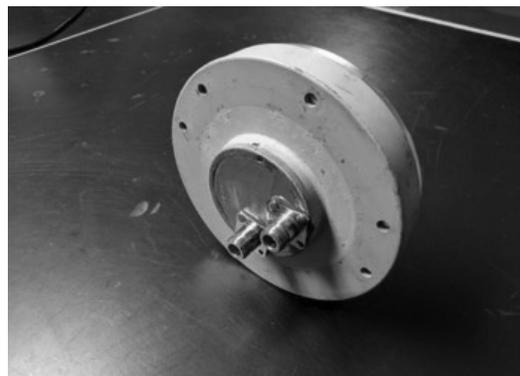
The completed “Chaparral” Patch

- Add driven patch and probes for V-pol & H-pol
- N-Type connectors for V-pol & H-pol
- Use plenty of JB Weld epoxy to hold it all together

Front View – Completed Patch



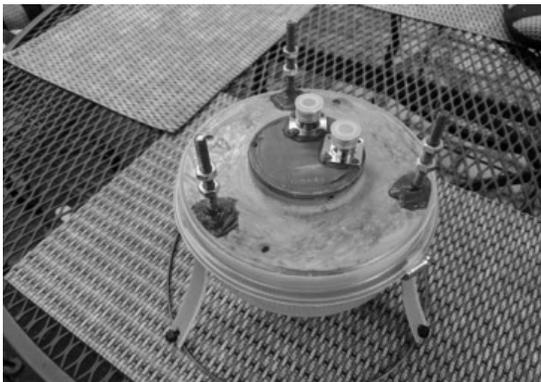
Rear View – V & H-Pol N-Type F



Full Assembled Patch Feed

- Add three brass feed mounting bolts
- Add Radom (Rubbermaid food container #3)
- Add BFR – Beam Forming Ring

23cm Band Patch Feed



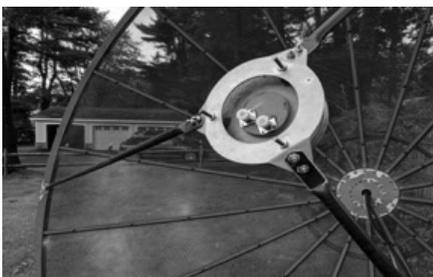
Installed in TVRO Dish



The “Chaparral” Patch mounting

- Feed position is adjustable with 3” bolts / nut position
- Add Radom (Rubbermaid food container #3)
- Beam Forming Ring supported on fiberglass strips

Feed Mount Bolts



Rubbermaid Radome



Adding the BFR



Adding Hybrid Divider

- CP is generated by feeding V-pol & H-pol – 90 Deg.
- Unused Hybrid port is terminated in 50 ohms
- Preamp is isolated from TX power with T/R relay



23cm Patch – On-Air Experience

- Sun Noise = 10 dB / ON0EME = -11 on WSJT
- On Air results in 2018
- 170 QSOs (147 - JT65 / 3 - CW)
- 107 Initials
- WAC
- VUCC - 80
- WAS - 22
- DXCC - 34

23cm Patch on Toki 550 Umbrella Dish

- My interest in the 23cm Patch was focused on the Toki
- I needed a small / light weight feed for a portable dish
- The patch and hybrid mount nicely on the button hook

23cm Patch on Toki 550



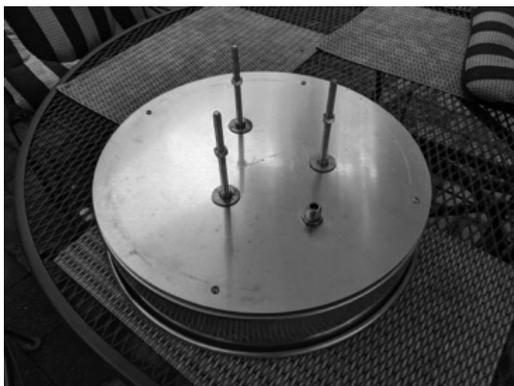
Hybrid & Radome on Patch



70cm EME with Cake Pan Feed

- The same feed mounting holds a 70cm loop feed
- Use a 16" x 3" Cake baking pan (aluminum)
- Loop Feed design by OK1DFC

Feed Mount Bolts



70cm loop feed mounted



70 cm Loop - On-Air Experience

- Sun Noise = 6 dB / Estimated 20 dBi gain
- On Air results in June 2018 / 180W SSPA
- 18 QSOs
- 14 Initials
- VUCC - 14
- WAC - 3
- WAS - 3

Credits – These are the Designers

- Thanks to SM6FHZ / SM6PGP for the 23cm Patch Feed design
- [http://moonbouncers.org/
Circularly%20polarized%20patch%20feed%20for%201296%20MHz
A.pdf](http://moonbouncers.org/Circularly%20polarized%20patch%20feed%20for%201296%20MHz%20A.pdf)
- Many thank to OK1DFC for the 70cm Loop Feed design
- <http://www.ok1dfc.com/eme/technic/432feed/432feed.htm>
- Thanks to KL6M for access to Septum feed parts
- Thanks to G4DML for practical patch experience sharing

Summary

- North America is blessed with an abundance of TVRO dishes
 - Using the factory feed support system can speed construction
 - Mount different feeds using a “standardized” support system
 - The 23cm Band Patch is worthy of a close look
 - The 70cm Loop Feed is easy to build and performs nicely.
 - Standard mounting allows 23cm to 70cm feed swap in 30 mins.
 - Future plans to build standard mount for Septum Feed.
-
- Thanks to K2UYH and all my other Elmers.

Band and PTT outputs using K3 RS232 signals

Dave Robinson WW2R



The most annoying "feature" of the Elecraft K3 is the lack of a separate PTT output for each band, especially when using it as a transverter driver for many microwave bands.. Previously I had used a PIC to decode the band data appearing on the K3 accessory socket pins. However every time I reconfigured the transverter "XV" assignment I needed to reprogram the PIC which was unacceptable. It was realised that the RS232 stream from the K3 showed the frequency is on irrespective of transverter assignment, so it was decided to use this information to drive the new unit. It would allow a "1 of 8" open collector output to switch the appropriate external device per microwave band and also a "1 of 8" open collector PTT output for the appropriate band. To keep an eye on the switch status a 16x2 LCD display would be used. To avoid issues with RS232 levels a true RS232 level converter (MAX232) was employed, despite the extra cost.

First question was what RS232 string to look for from the K3? In the manuals I found a config:autoinf =1 which causes the k3 to send an IF sentence out every time the frequency changes. This change can be caused by tuning the radio with vfo or pressing band up or down. The IF sentence gives the following into (from k3 programmers guide)

IF (Transceiver Information; GET only)

RSP format: IF[f]*****+yyyyrx*00tmvspbd1*; where the fields are defined as follows:

[f] Operating frequency, excluding any RIT/XIT offset (11 digits; see FA command format)

* represents a space (BLANK, or ASCII 0x20)

+ either "+" or "-" (sign of RIT/XIT offset)

yyyy RIT/XIT offset in Hz (range is -9999 to +9999 Hz when computer-controlled)

r 1 if RIT is on, 0 if off

x 1 if XIT is on, 0 if off

t 1 if the K3 is in transmit mode, 0 if receive etc

Naively as the “t” character reflects Receive or Transmit state I assumed the IF sentence would be sent when changing RX-TX or TX-RX **WRONG.. it doesnt!** I asked Wayne N6KR if the transition between transmit and receive state could trigger an output but he said the RS232 port was already busy enough! So as well as looking at the RS232 port I would have to take the PTT signal out of the k3 and monitor it as well

Thoughts then turned to which PIC, A PIC with more pins than the previous 16F628 would be needed, the 16F886 was chosen as it had been used in other amplifier monitoring projects. It turned out that even this did not have enough output pins for what I needed so a high current SPI interface shift register was used (TPIC6B595 as used in the Elecraft KRC2) for the band output signal and the directly driven (quicker) ULN2803 for the band specific PTT. The bands decoded were the USA VHF/UHF/Microwave bands. The initial circuit was built on a PCB produced by express PCB and tried.

Initial testing was good, band outputs switched as designed but the PTT output wasn't behaving sensibly. It was soon realised that if the PIC was looking all the time for RS232 data coming in then it could not also be looking at the PTT signal. Time to get acquainted with external Interrupts (not used since my GPO introduction microcontrollers class in 1984!). Unfortunately the external interrupt pin (21) on the 16F886 was already used for the LCD port so some pin juggling (and a new PCB) would be needed. The final circuit is as follows:-

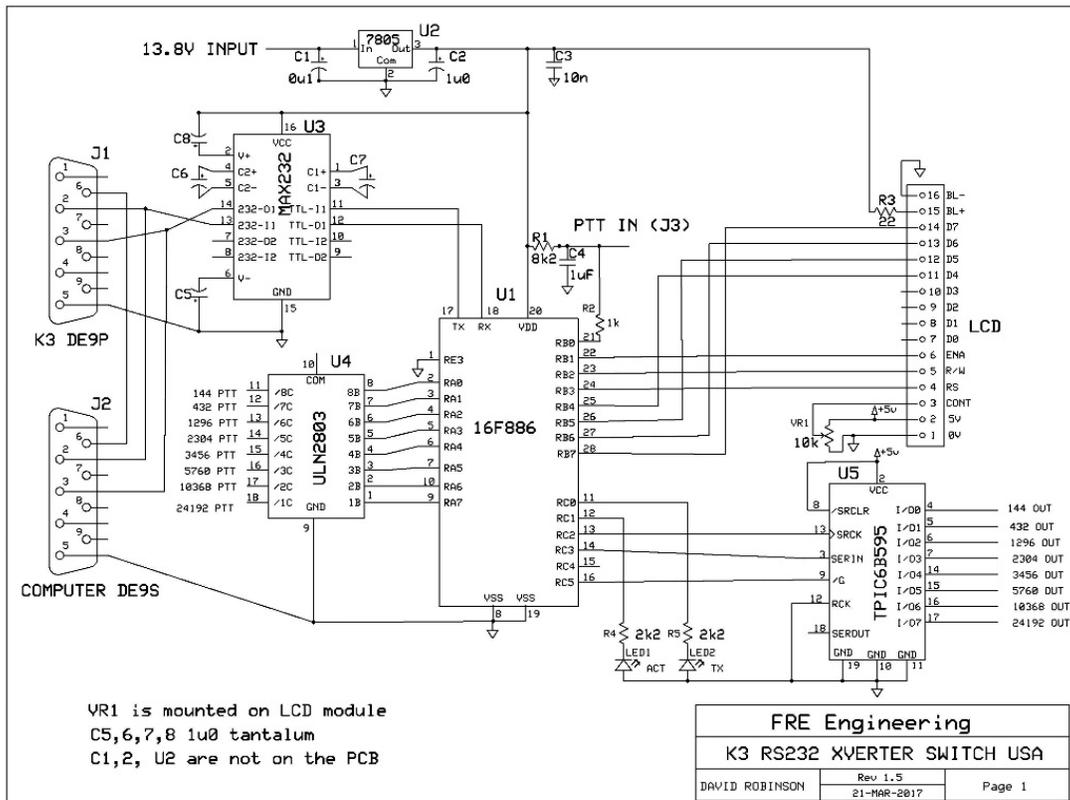


Figure 1 Circuit diagram

Circuit

The PIC monitors the RS232 port and uses the data to send the appropriate band through the PIC's SPI port to the TPIC6B595. If the PTT input signal goes low (ie the rig is on Transmit) an interrupt is generated that causes the PIC to switch the band applicable ptt open collector output. During this period, it ignores what is happening on the serial port so if you move frequency while you are on Transmit (very unsociable) or hit the band up button on transmit (hope you have a good sequencer!) the band will not be updated. Debounce circuitry (R1/R2/C4) is applied to the PTT port. With Pin 1 earthed the unit expects 38400 baud data from the K3. If it is connected to +5V the unit expects 9600 Baud data from the K3

Construction

A small 4" x3" PCB was designed and the original PCB produced by Expresspcb. However the design has been ported to Robot room copper connection software allowing gerbers to be produced and the PCB made by OSHPark in a stunning purple! The circuitry is boxed in a 5x3x2" aluminium box. The front panel contains the LCD, activity LED and Transmit LED. The rear panel has two 9 pin connectors DE9P and DE9S for RS232 in and out (to allow the decoder to be used inline with other devices), a DE25S 25 pin connector for the decoded outputs, an RCA connector for PTT input and a 12 connector

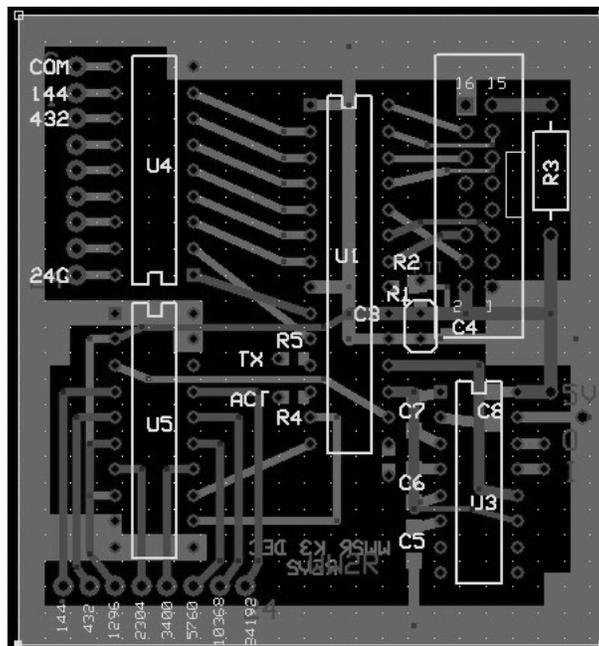


Figure 2 PCB Overlay

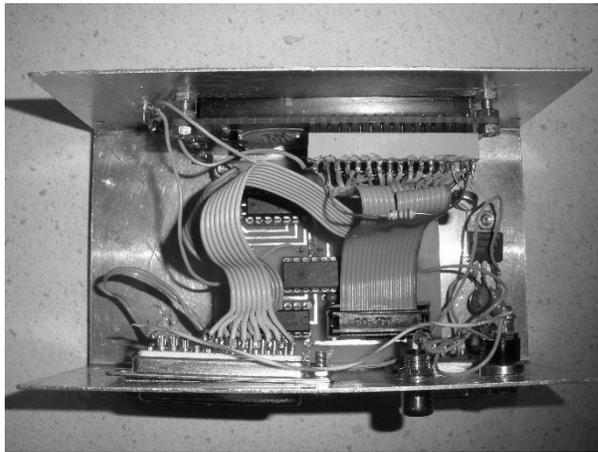


Figure 3 Internal view of the decoder

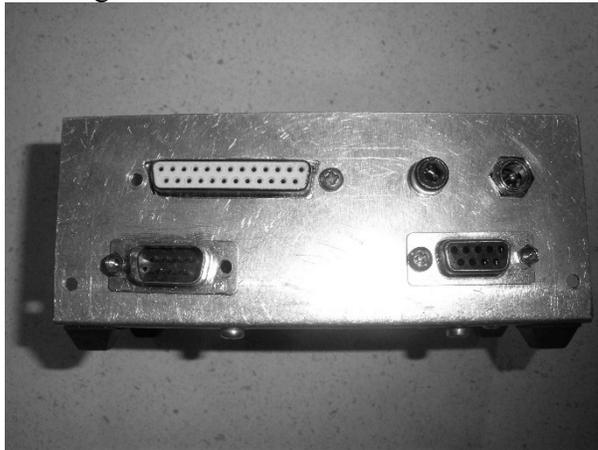


Figure 4 Rear Panel view of the decoder



Figure 5. The decoder in use on 6cm



Figure 6. The decoder in use on 9cm

To test the outputs I connected two ten segment LEDs to the two sets of outputs with the common of each connected through a 2k2 resistor to 12V. A segment would illuminate when the output goes low. The following is what happens on 6cm:-

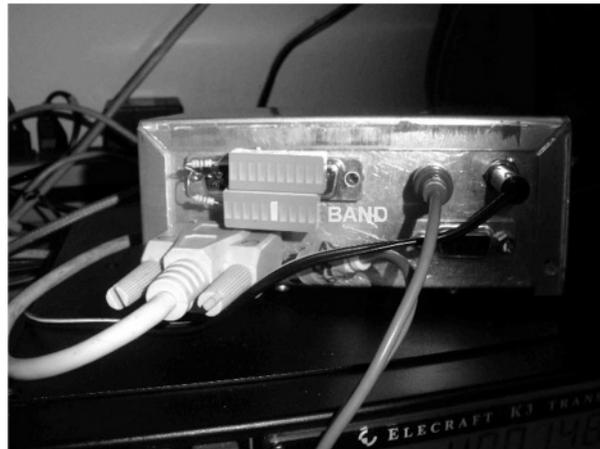


Figure 7 Receiving on 6cm

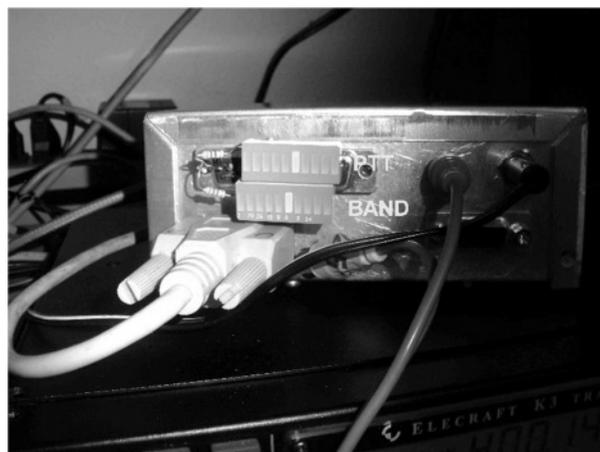


Figure 8 Transmitting on 6cm

Deployment

In the K3 menus

Set CONFIG:AUTOINF to 1

Make sure CONFIG: RS232 is set to 38400

Connect the RS232 cable coming out of the K3 that would normally go to the computer to J1

Connect the PTT out from the K3 to the PTT Input of the decoder (J3)

Connect 8-14 volts to the unit

Turn on the K3. it should produce an IF sentence and update the frequency on the display. If not move the VFO knob, even 1Hz which will cause the display to update

If you have setup transverters for 2m (internal or external; it does not matter) 70cm 23cm 13cm 9cm 6cm 3cm and 1.2cm as you step through them the band output will conduct to ground. On Transmit only the specific band ptt output will conduct to ground. It does not matter which transverter is set up in the K3 menus as XV1, XV2 XV3 through XV9 as only the operating frequency is read

Anything below 100MHz is considered HF and no output is provided. the LCD will however show "HF"

Note that the actual frequencies designated to be on each band are not overly specific. Example 13cm can be 2301, 2304, 2310, 2320 2402 or 2424MHz. The decision is made on whether "Units of GHz=2". Similarly 24G can be 24048 or 24192. The decision is made on whether "Tens of GHz=2"

Conclusion

Objective met for VHF/UHF/microwaves. The design also works with the KX3.



An introduction to Oscar100



Noel Matthews – G8GTZ



Topics



- Oscar 100 overview
- Why is it a game-changer?
- What does it offer?
- How do I get started?
- Narrow band operation
- Wide band operation
- The WebSDR





What is Oscar 100

- Oscar 100 is 2 amateur radio transponders hosted on the Es'hail-2 Direct Broadcast TV satellite
- Owned by Es'hailSat in Qatar.
- Built by Mitsubishi Electric Company (MELCO) in Japan.
- Collaborative project with Es'hailSat / AMSAT-DL / Qatar ARS
- The first ever amateur payload on a commercial geostationary satellite



Oscar 100

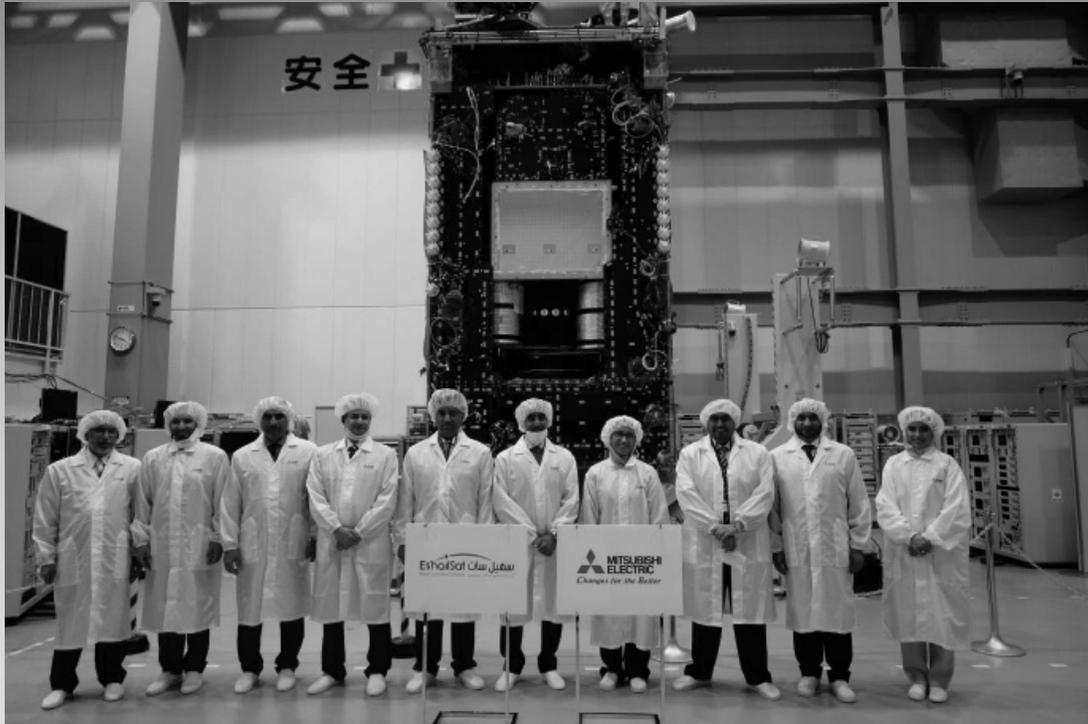
- Project started in 2012 by Qatar Amateur Radio Society and AMSAT DL
- Launched by SpaceX Falcon 9 from Cape Canaveral – November 2018
- Commissioned and ready for use in February 2019



Es'hail (Canopus) is the name of a star which becomes visible in the night sky of the Middle East as summer turns to autumn.



Es'hail-2

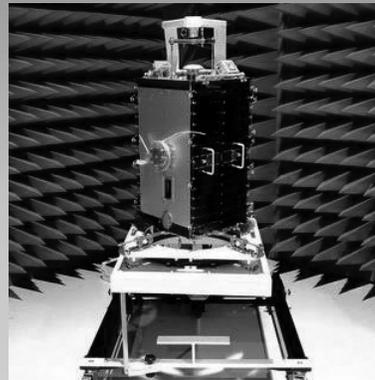
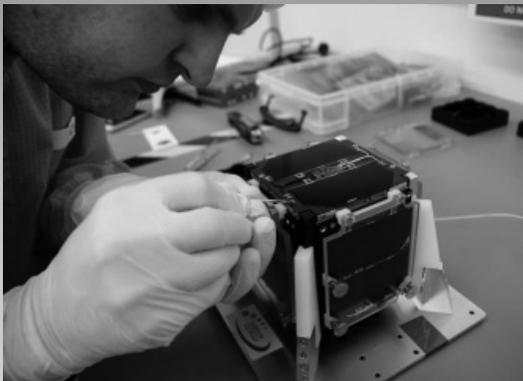


"Normal" amateur satellites



FUNcube-1 CubeSat AO-73

European Student Earth Orbiter (ESEO)



Based on a 10cm x 10cm x 10cm format.
- approximately 900g

MicroSat - 50kg

Orbits and coverage

- ☞ Low Earth Orbit
 - Typically 400 – 700km altitude
 - Orbit once every 90 minutes = tracking



- ☞ Medium Earth Orbit
 - 8000km - 20,000km
 - Used by navigation satellites
 - No amateur satellites

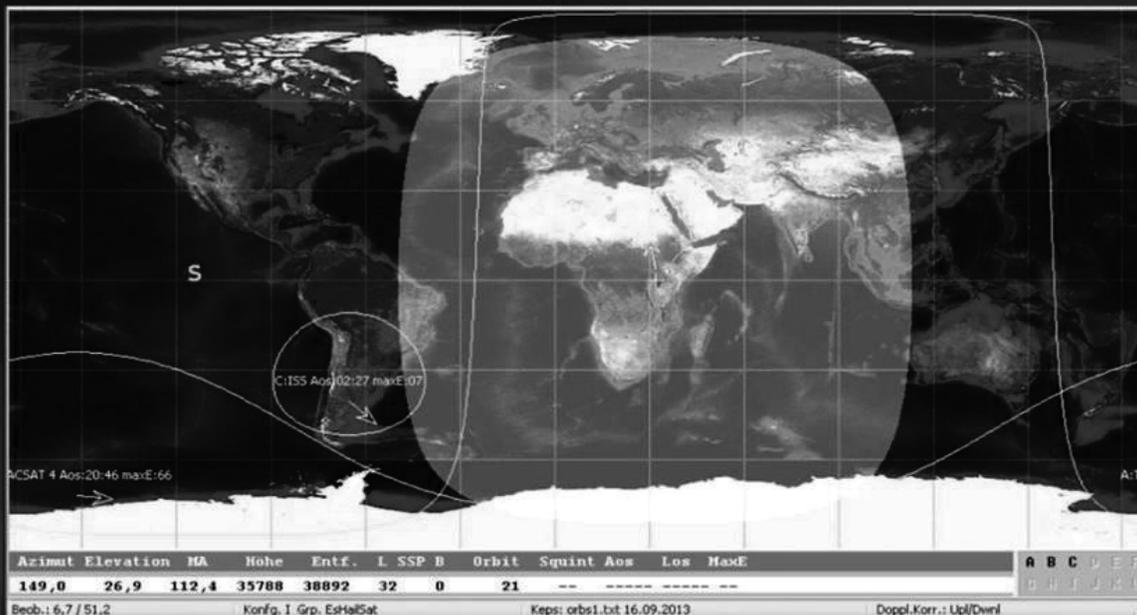
- ☞ Geostationary
 - 36,000km altitude
 - large coverage area – 40% of the earth and 60% of population
 - No antenna tracking needed
 - Where all broadcast TV satellites are

36,000 Km altitude

-3dB Beamwidth = 17.4° → ~20dB Antenna Gain !!



Earth Coverage Es'HailSat-2



What is on Oscar100?



- E 2 transponders dedicated to Amateur Radio
 - 13cms (2400MHz) uplink
 - 3cms (10GHz) downlink

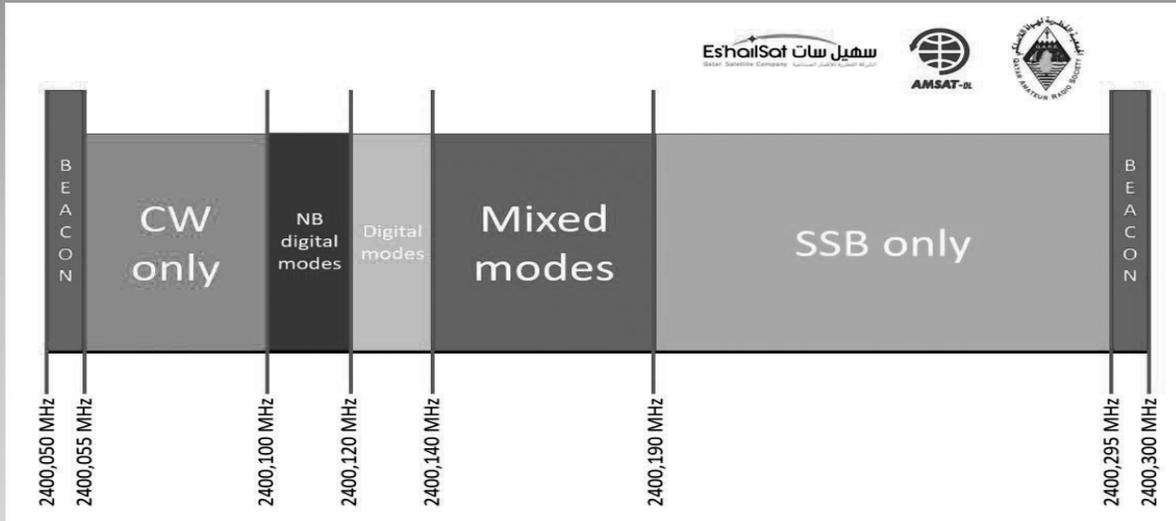
- E Narrow band transponder 250kHz wide
 - CW, SSB data modes etc
 - AGC and Leila over power warning system
 - CW and BPSK beacons

- E Wide band transponder 8MHz wide
 - Dedicated to Digital modes
 - Primarily Digital Amateur Television
 - Up to 8 DATV signals simultaneously
 - HD beacon channel





Narrowband Band plan



Wideband band plan



Beacon and Simplex DATV				Simplex DATV				RB-TV			
Simplex 1MS				Simplex 1MS				125KS			
Narrow Beacon								125KS			
Simplex 2MS				Simplex 2MS				333KS			
Wide Beacon				Maintenance Uplink				333KS			
2401.5	2402.5	2403.5	2404.5	2405.5	2406.5	2407.5	2408.5	2409.5			
Uplink (MHz)											
10491.0	10492.0	10493.0	10494.0	10495.0	10496.0	10497.0	10498.0	10499.0			
Downlink (MHz)											



How do I get started?



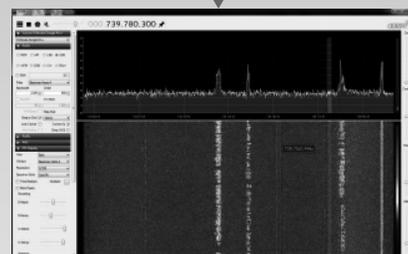
- Whether going for Narrow band or Wide Band DATV – start with receive...
- Satellite dish pointing at 26 degrees
 - 60cms (Sky) for NB
 - 90cm - 1.2m for DATV
- <https://eshail.batc.org.uk/point/>
 - Just south of Sky/Freeview
- Use a new PLL LNB for greater stability
 - Available for approximately £10



Is it this simple?

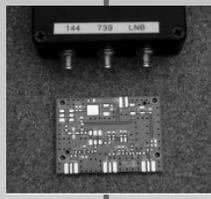


- Yes!
- A simple NB rx system is:
 - Sky dish
 - New PLL LNB
 - ~ £10 RTL dongle or Funcube, LimeSDR or Pluto
- Bias Tee to supply 12v
- Free SDR software
 - SDR#
 - SDR Console
- Tune to the IF frequency of 739 MHz



Can I use a VHF / UHF rig?

- Yes – but...
- The output from the LNB is 739MHz
- A downconverter will shift this to 432 or 144MHz
- Frequency stability is an issue
 - Lock all oscillators to external ref
 - Use SDR locking



NB Transmitting - 1

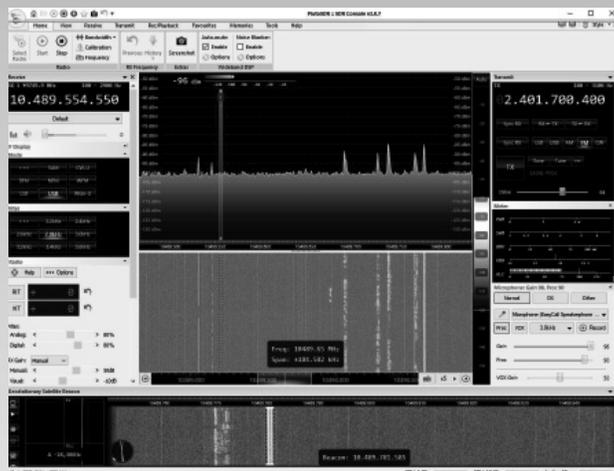
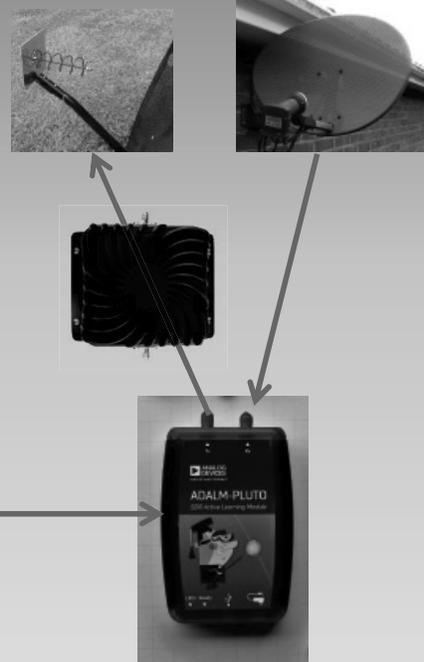
- The NB transponder is VERY sensitive
- Transvert up from a VHF or UHF rig
- Small PA ~ 4 watts
 - wi-fi booster
- LHCP helix dish feed
- Separate dish or dual band patch feed





NB transmitting - 2

- SDRconsole by G4ELI
- Tx and Rx via Pluto or LimeSDR
 - Full duplex
 - Frequency lock to BPSK beacon



NB operation

- All modes permitted
- Digital, SSB, CW, Hellschreiber...
- Great for experimentation and easy to receive
- DL7NX
 - 1 watt to 4 ele PCB Yagi.



- PA3WEG
 - 1 watt to a PCB quad patch



Oscar 100 Wideband



- Oscar 100 wideband is an “8 MHz bent pipe” transponder for wideband digital use
- Occupied bandwidths can be 200 kHz – 8 MHz
- Most signals are <1MHz wide
- Some experiments below 100Khz
- DVB-S2 with H264 / H265 video



Receiving DATV



- Downlink frequency is 10,491 – 10,499 MHz and within pass band of standard consumer LNB
- PLL LNBS should be used to give stability for Reduced Bandwidth TV signals
 - Locking can cause phase noise problems
- However 9,750 MHz LO puts IF outside consumer set top box tuning
- 90% of signals are Reduced Bandwidth (RB-TV) and cannot be received on a consumer STB



MiniTiouner USB tuner



- ☞ A wide frequency range tuner
 - Covers 143 – 2450 including 741 MHz
- ☞ Available as kit or built unit
- ☞ PC based with software by F6DZP
 - Gives totally flexible receive system
 - MPEG-2, H264 and H265
 - 33Ks to 27 Msymbols DVB-S, DVB-S2, for HD-TV, DATV and RB-TV

☞ See <https://batc.org.uk/>



Receiving DATV



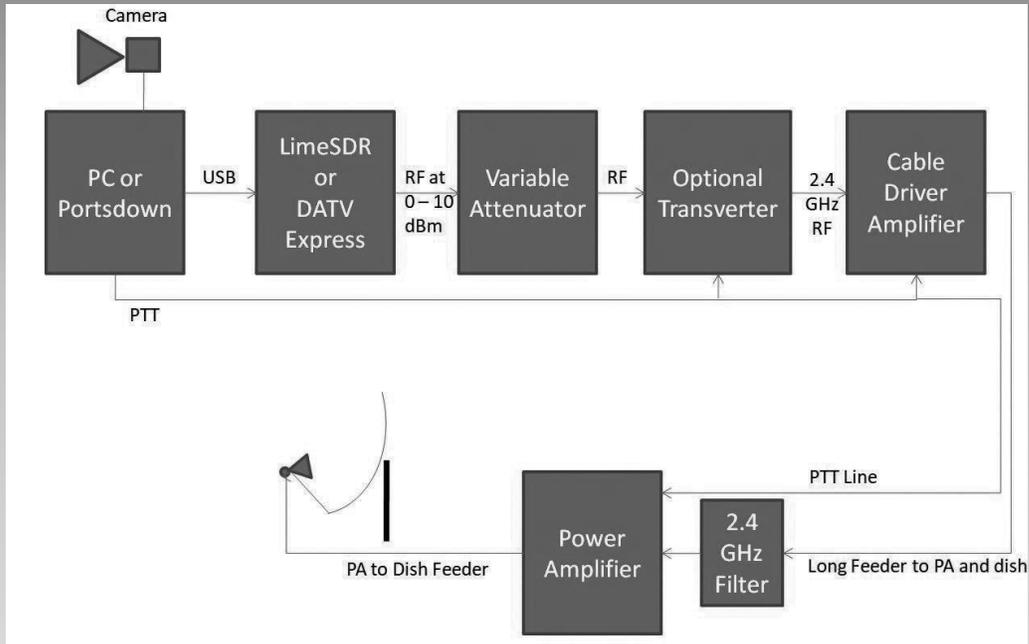
- ☞ Aim for a 1m dish
- ☞ Check your dish direction using
 - <https://eshail.batc.org.uk/point/>
- ☞ Align using BADR-4 TV services
 - 12,597 MHz, 27500 Ms, Horizontal
 - ~11dB MER
- ☞ Check the WB beacon
 - 2Ms DVB-S2



☞ More details:
https://wiki.batc.org.uk/Receiving_Oscar_100_DATV_signals

Dish size	Received MER
1.8m	10dB
1.2m	8dB
1m	6dB
80cm	5dB

DATV transmit system



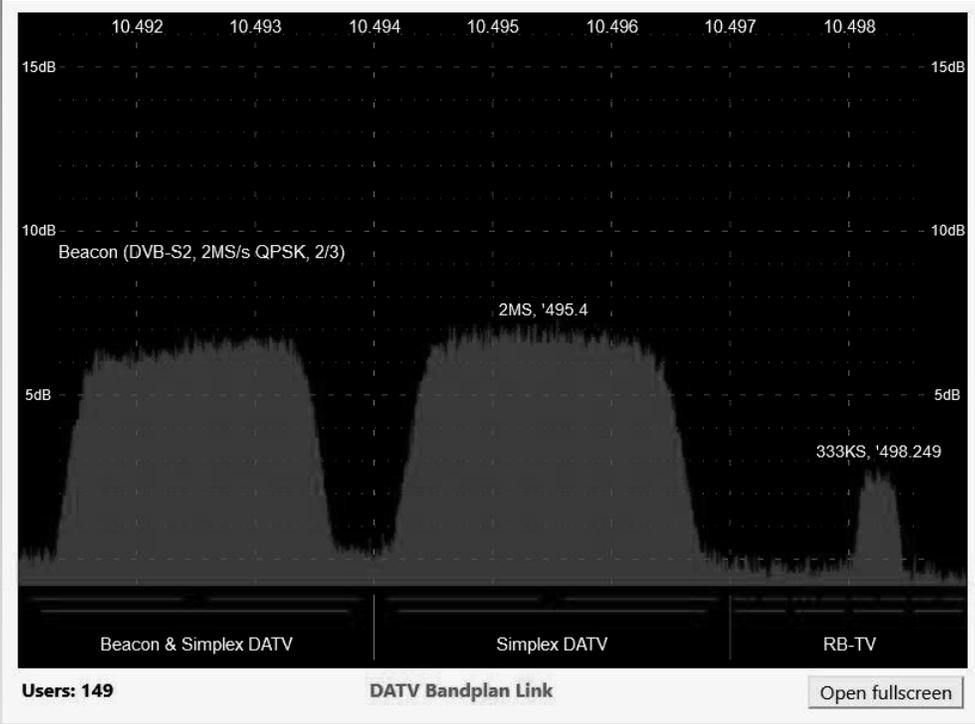
DATV transmit

- ~30 watts in to a 1.2m dish
- PA at dish and VERY short feeder
- Dual band dish feed
 - 2.4GHz patch
 - LNB 22mm waveguide





3 signals

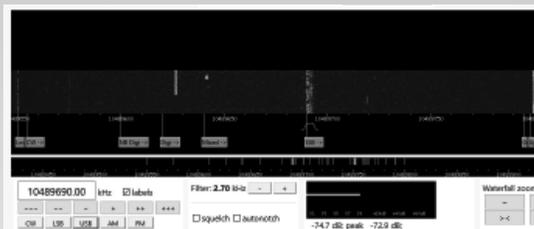


F4HSL ~ 80KHz!



The WebSDR

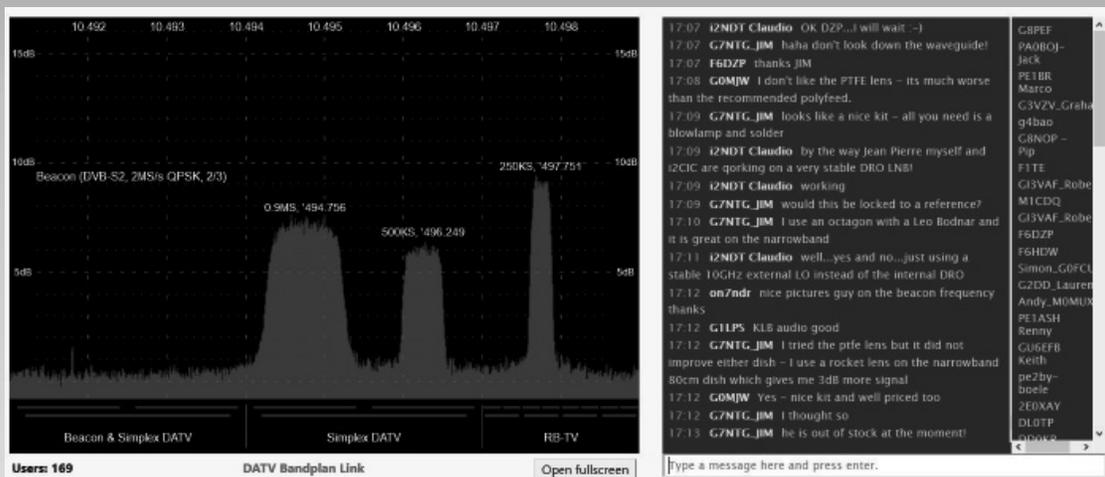
- AMSAT-UK and BATC wanted to make Oscar 100 accessible to everyone
- An on-line WebSDR which only needs a standard web browser
- Full coverage of NB transponder with waterfall and full audio decode.
- 350+ users on first weekend





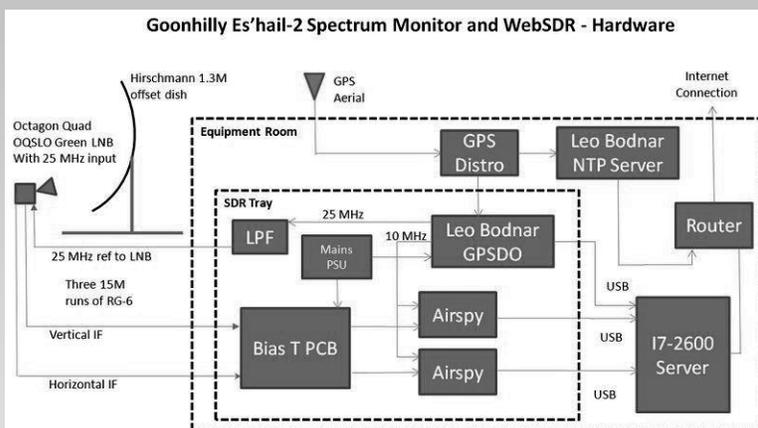
Spectrum Monitor

An essential tool to enable the Wide Band transponder usage



BATC + AMSAT-UK SDR

- Located at Goonhilly Earth Station
 - Quiet secure location (IO70JB)
 - Excellent network connectivity
 - Scaled for 500+ users





Is it really amateur radio?



- Absolutely - hundreds of people are engaged in that most vital aspect of amateur radio:
 - Self training in wireless telegraphy
- It has breathed new life in to the satellite and microwave communities
- As well as providing 24/7 communications to 1/3rd of the earth



Conclusions



- Oscar 100 is a fantastic opportunity for amateur experimentation
- Receive is easy!
- A good transmit capability is more of a challenge but not impossible!!
- Start simple
 - Get a receiver working!





WebSDR demo

- Usable by anyone with a web browser
 - Scaled to support 500 simultaneous users
- All listening to different frequencies and decoding different modes!
- Runs s/w developed by www.websdr.org
 - More than 150 systems around the world
- <https://eshail.batc.org.uk/nb/>
- Wideband spectrum monitor
 - <https://eshail.batc.org.uk/wb/>

Microwave Update 2018 Antenna Range Results

Frequency	Call	Description	HB/Com	Measured	Gain dBi	Comments
1269 MHz	WA5VJB	Reference Horn	Com		12.0	FAA L-Band Horn
1269 MHz	WA5VJB	11 ele PCB Yagi	HB	-0.8 dB	11.2 dBi	Prototype for Arrow
"	WA5VJB	11 ele PCB Yagi	HB	-0.7 dB	11.3 dBi	#2 Prototype for Arrow
1296 MHz	W1AIM	Yagi	HB	.8 dB	12.8 dBi	10 over 10 Skelton Slot K2VXO built
	VE2UG	WG Slot	HB	-3.4 dB	8.6 dBi	4 Slot on Axis
	W1AIM	12 turn Helix	HB	-2.2 dB	9.8 dBi	4 Slot on edge
				.8 dB	12.8 dBic	Axial Ratio 2.6 dB
2304 MHz	WA5VJB	Reference Horn	Com		17.0 dBi	Scientific Atlanta
	WA8OGS	2.4 WiFi	Com	-2.0 dB	15.0 dBi	Patch array
	WA8OGS	2.4 WiFi	Com	Interference	---	Colinear type
2425 MHz	WA8OGS	2.4 WiFi	Com	-1.4 dB	15.6 dBi	Patch array
5760 MHz	WA5VJB	Reference Horn	Com		22.2 dBi	
	W5VY	MA-WA58-1X PM	Com	-4 dB	21.8 dBi	
	K2EZ	Dual Band Feed	Com	6.8 dB	29.0 dBi	24" dish DEMI Feed
	K5VY	Dual Band Feed	Com	8.0 dB	30.2 dBi	24" dish DEMI Feed
10 GHz	WA5VJB	Reference Horn	Com		20.5 dBi	
	W7TXT	6 Slot	HB	-9 to -14	11.5 dBi on Axis	
	W5VY	Dual Band Feed	Com	11.5 dB	31.9 dBi	24" dish Demi Feed
	K2EZ	Dual Band Feed	Com	13.1 dB	33.6 dBi	24" dish Demi Feed
24 GHz	WA5VJB	MA Reference Horn	Com		21.4 dBi	
	W1GHZ	Horn	HB	-2.6	19.1 dBi	"Maker" machined 2.4x4.5x4.6 cm
	VE2UG	Patch array	Com/HB	7.6	29.3 dBi	Modified 256 ele Patch array



Elcom CDFSL Synthesisers

by Roger Ray, G8CUB

Local Oscillators for 10GHz upwards 9936 MHz, 10224 MHz, 12240 MHz etc. I purchased a pair of Elcom ILCDFSL-1295 synthesisers from 'art-in-part' on ebay. These were intended for a millimetre band project. However once I started looking at them, there potential use for 10 & 24GHz became apparent.

They use just a 12 and 8V supply, have a 10MHz reference and can be set to any frequency that is a multiple of 10/3 MHz. The ones purchased were described as new, they certainly had seals intact, and included a test report. On connecting the 12 & 8V supplies, the first unit came up on 12.6 GHz. The second unit for some reason was damp, when trying it, the supply currents fluctuated wildly, and there was no output. It was for this reason I unscrewed the multiple cover screws and dried it out. The was to prove the key to it's use at 10GHz. Because while I was checking it's operation, I discovered that the VCO was running at one fifth of the output frequency.

The normal output frequency range was 12.2 – 12.95 GHz. Would it be possible to use it on 4/5 frequency? A check with a calculator, showed that the sums worked. If set to 12420MHz, 4/5 frequency is 9936MHz (10368-432), if set to 12780MHz 4/5 frequency is 10224MHz (10368-144).

Initially I had not tried programming the unit, so I used the default 12.600GHz, and looked to see if I could get output at 10.080GHz. Internally there is some serious filtering to select the 5th harmonic of the VCO, so this would need re-tuning.

I tried the piece of ceramic ‘trick’ that I used on the Ceragon modules for 5.7GHz. With a little bit of size trimming, it worked more or less straight away!

The idea of using these modules came from the excellent article by Dave G4FRE WW2R. Dave had done the hard part in working out the programming, and then writing a PIC program that allows the frequency to be set.

I must admit I struggle with the digital stuff. This was one of the reasons that I was keen on the Alcatel synthesisers, that did not need any programming! My initial attempts to program a PIC went down the usual course for me, of being a failure. I bought a cheap JDM programmer from ebay. I then found that I could not use it with an USB / RS232 adaptor. When I eventually got an old PC with dedicated RS232 working, I found that the software did not work either. With advice from Dave, I got hold of a PIC Start + programmer, but that needed upgrading to work with the 12F675 PIC.

So what finally worked for me was the JDM programmer, a 1:1 RS232 lead, win pic programmer software, and MPLAB_IDE software to compile the hex code (both free on the web). Although later on I found the upgraded Pic Start + easier.

Now I could look at setting the frequency that I wanted. Following Dave’s article, I set the frequency to 12.420 GHz. I edited the .asm program using Word Pad, putting in 37 260 (12,420 x 3). Programming the PIC gave the correct output frequency, then dropping on the pieces of ceramic, and moving them around with a trimming tool, I had 9936MHz at +12dBm output. The plots below show that both the close in and wideband spurious were very good.

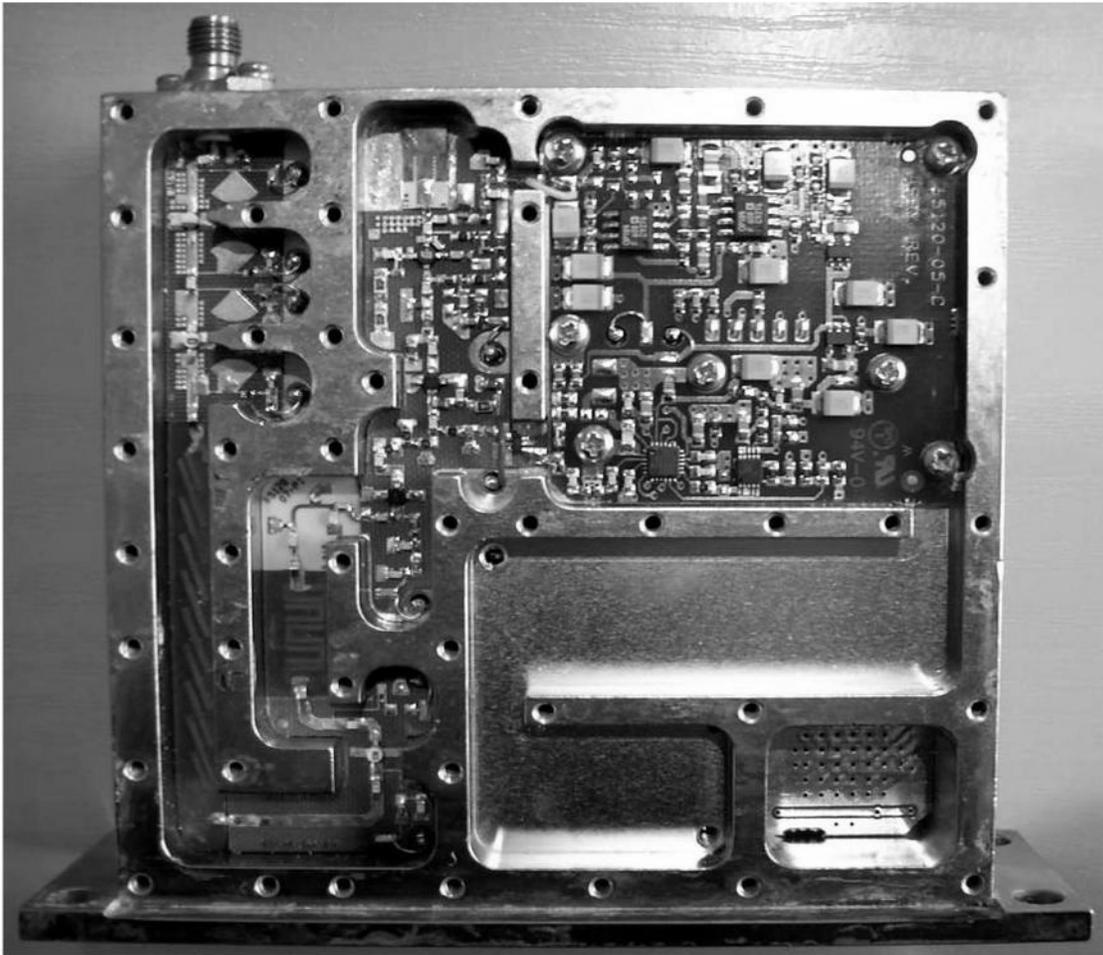
Using the KE5FX software to make a phase noise plot gave an excellent result, some 20dB better than the original Alcatel synthesiser, when compared at final frequency. The result of 81.6dB at 1kHz offset, would require >122dB at around 100MHz in a oscillator multiplier combination (phase noise increases 20dB/decade). The plot was taken using the internal 10MHz reference. Surprisingly this reference looks adequate for 10GHz use, once set on frequency! Although this was tricky, as it really needs a small square trimming tool. Frequency shift against time was measured as 1.2kHz for the first 30 minutes, and less than 100Hz over the next 2 hours. The addition of a miniature relay driven from one of the unused connector pins, would make a neat system to later add an external 10MHz / GPS locked reference if desired. Further operational use has shown the internal reference to be very good both noise and stability wise. It is useable after a minute or two warm up!

Supply requirements:

+8V 260mA typ. (+7V min.)

+12V 120mA typ. (+11V min. (+12V for optimum performance))

Pin-out is as per the references - 8 way in line Molex connector, or 15 pin D-type

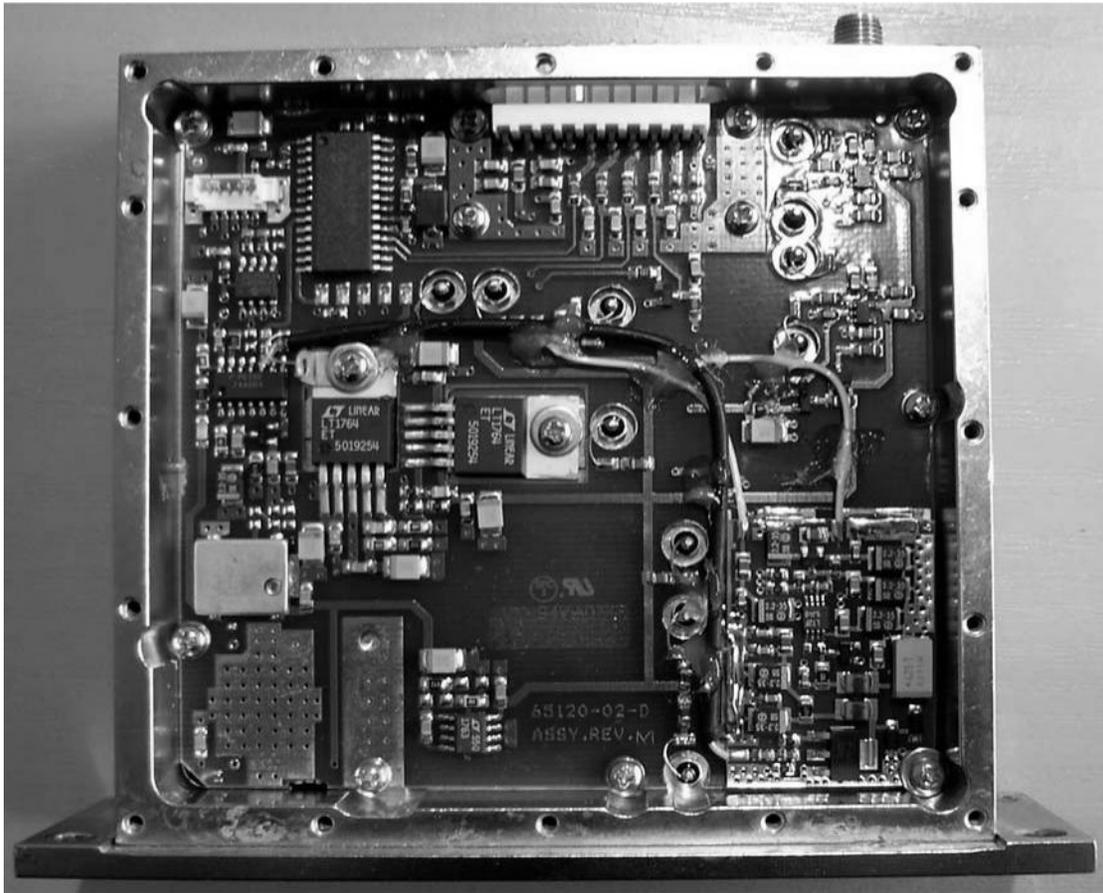


RF side of the synthesiser showing the hairpin and stripline filters at the bottom left. Some useful frequencies CDFSL-1201 (11.2 – 12.0GHz)

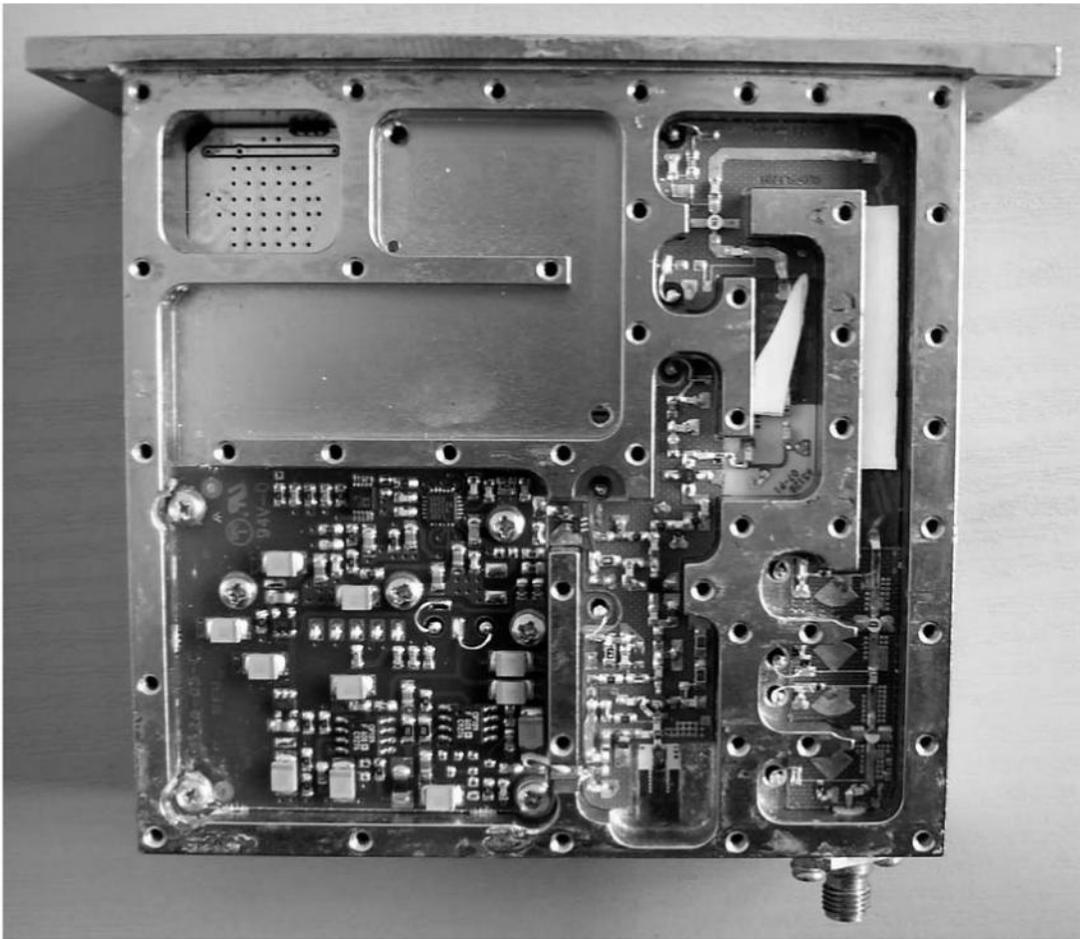
Frequency	Freq. set	Modify	Code (fx3)	Use
11880 MHz	11 880		35 640	47088+432 / 4

Some useful frequencies CDFSL-1295 (12.2 – 12.95GHz)

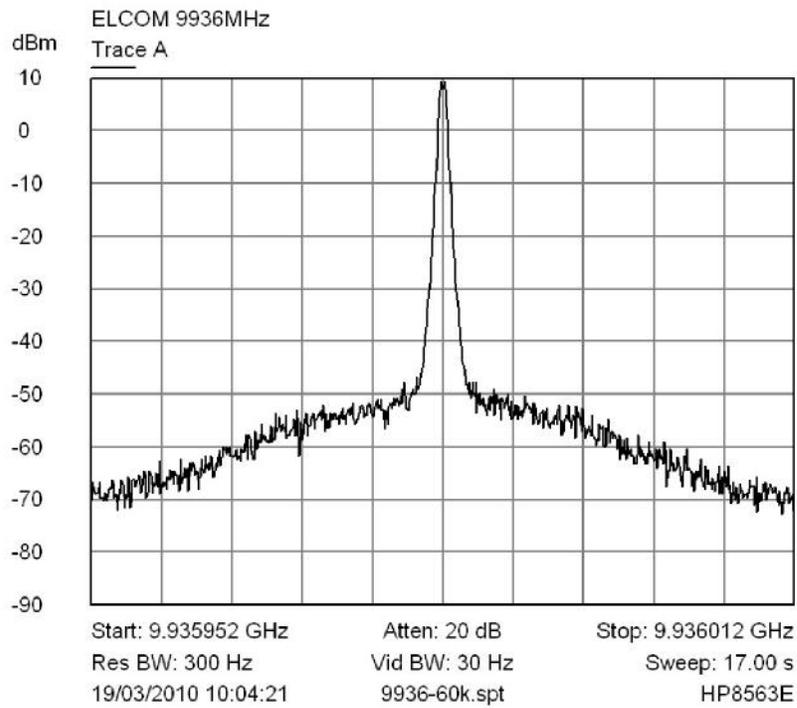
12240 MHz	12 240		36 720	24048+432 / 2
9936 MHz	12 420	x	37 260	10368 - 432
10224 MHz	12 780	x	38 340	10368 - 144



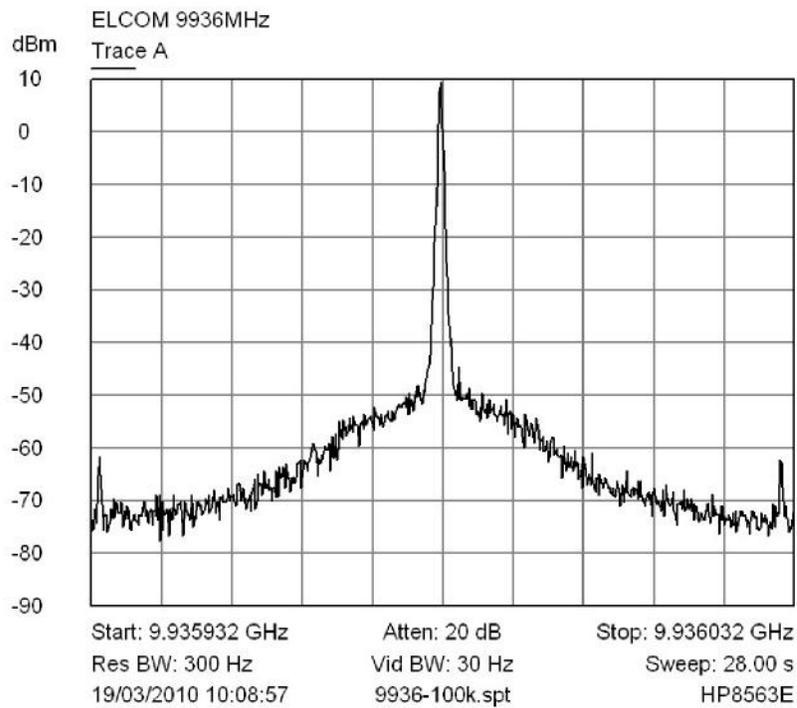
Reference / DC side of the synthesiser. The track at the extreme left of the reference TCXO upward carries the 10MHz signal. Connect into the capacitor at the top of the track when using an external 10MHz reference. Removing the +5V supply to the internal TCXO at the same time is probably a good move.



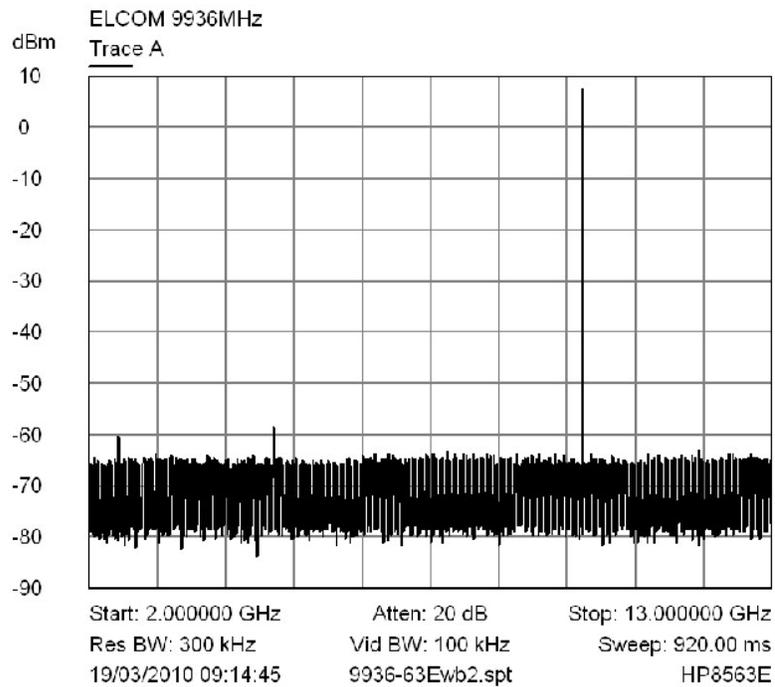
Modified RF side – ceramic in place on the filters. After this picture was taken, the ceramic was glued in place with araldite.



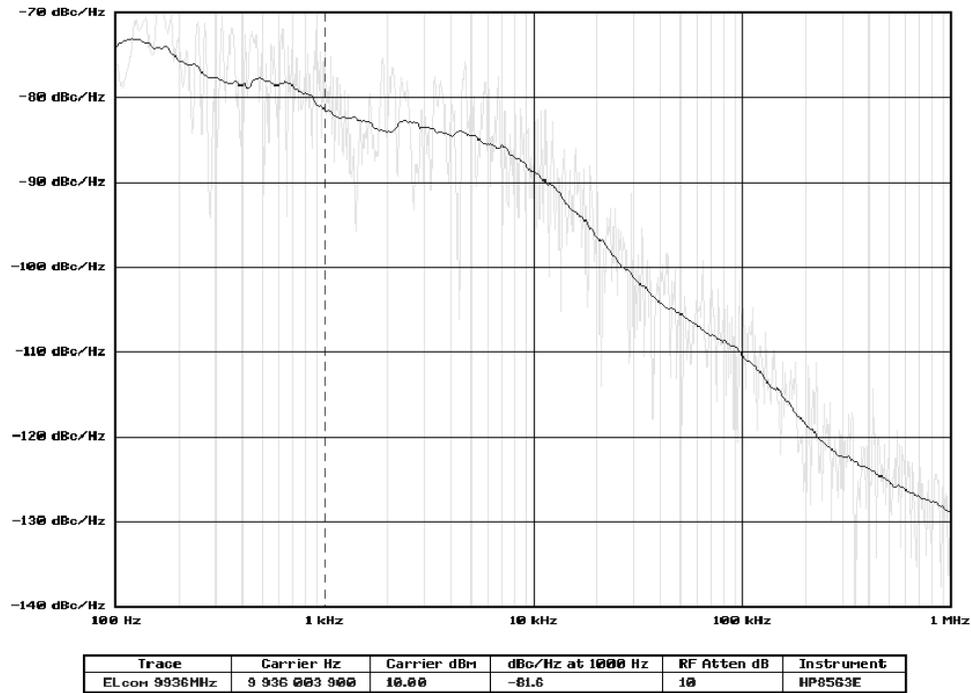
Spectrum Plot with 60kHz Span



Spectrum Plot with 100kHz Span



Wideband spectrum plot 2 – 13GHz



Phase Noise plot using KE5FX program

References: <http://g4fre.com/dfs1201.htm>

Note: When using the older DFS 1201 or DFS 1301 units, the regulated supply should be 6.0V not 8.0V. Also the programming on the 1301 is x2, ie 5MHz steps!

A look at the phase noise performance of a range of synthesisers and references

Part 1.



The move away from the traditional crystal oscillator / multiplier, to frequency synthesisers, has not been all beneficial. The improvement in getting directly to the required LO frequency, and stability when a good reference is used, has been offset by poorer, and often much poorer noise performance. Now a range of very good synthesiser chips have become available, from the likes of XXXX.

Although the chips are capable of good performance, to obtain excellent performance required great attention to detail. The good news is that low cost boards and complete units are available on ebay starting at around £15. There is a whole range of synthesisers with maximum frequencies of 4.4, 6.8 & 13.6GHz. The higher frequency units being particularly attractive to multiplying up for the millimeter bands. If, and that's a big if, the noise performance is good enough.

One of the big problems with wideband synthesisers is the gain of the VCO (sensitivity). If for example the VCO covers 6GHz with 6V, then the gain is 1000MHz / volt. So just 10uV of noise on the control line will FM modulate the VCO 10kHz. Thus ultra-low noise voltage regulators are required. This together with non-optimal layout and grounding, is where many of the low cost ebay synthesisers fall down.

The other problem is the accuracy, stability and noise performance of the reference. To obtain the best noise performance from the synthesiser, requires the use of high frequency references. The on-board references are not good enough for the higher microwave bands, or digital modes on the lower bands. The use of high performance external references is a way forward. Although some of the on board oscillators are on 'odd' frequencies. A good low phase noise reference should give the performance that we require inside the loop bandwidth of the synthesiser. This is often the only region we are really interested in. However when the noise is already on the control line, rather than within the oscillator, the best reference in the world, is not going to give the required noise performance.

It is intended to look at a number of commercially available synthesisers, and reference oscillators, and compare their noise performance. At one end of the cost range being a £15 ebay board, with an ERA 16GHz crowd funded signal generator at the other end at \$750 (when available). Somewhere in the

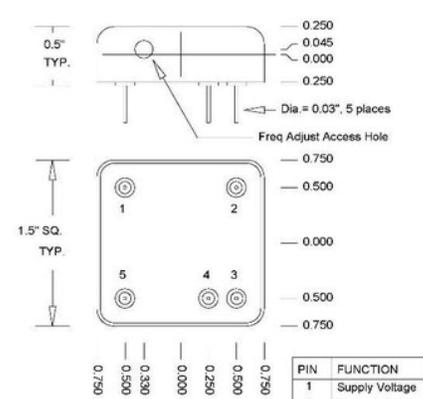
middle being the ZL 14G board, and the Kuhne 13.6GHz synthesiser. For the higher bands in particular what is required is a replacement for the much used Elcom synthesisers, which have become scarce. Preferably with the option of external reference and more frequency setting versatility, and lower noise.

For the really high bands above 76GHz, good phase noise of the LO becomes difficult to obtain. We are fighting against the increase of noise with multiplication of $20 \log n$. Where n is the multiple in frequency. Thus without any additional noise contribution noise at 10GHz will be 60dB higher than 10MHz, and 80dB higher at 100GHz.

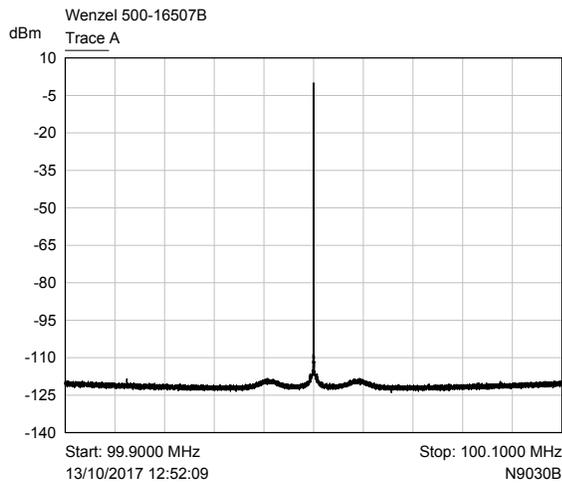
If starting at 10MHz an ultra-low noise reference is required. Talking with Brian Justin WA3ZMS, who holds many of the American firsts and records for the millimetre bands. He said he purchased two 10MHz references to his specification very early on, at an equivalent cost of a quality HF transceiver and linear!

A number of manufactures make low phase noise reference oscillators. Wenzel in particular make some superb units, but at a cost.

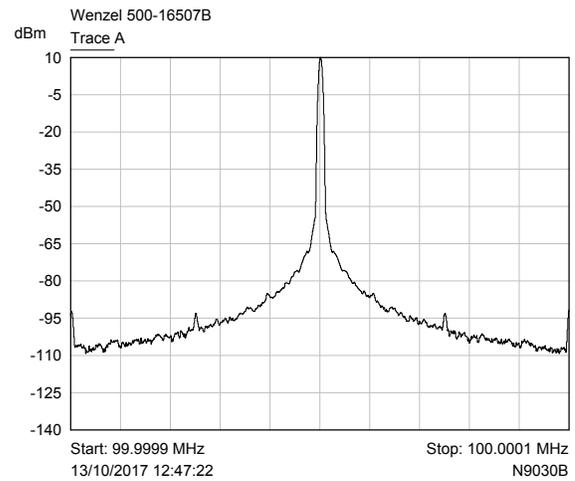
Recently a small quantity of Wenzel 100MHz units appeared on ebay from Israel. The noise performance is very good, using an SC cut crystal. Stability is not as good as a 10MHz oscillator, but may well be good enough for many applications. For ultra-stability, phase locking to 10MHz with a low loop bandwidth, may be required.

<p>OUTPUT Frequency 100 MHz Level +10 dBm ± 2 dBm into 50 ohms STABILITY Aging $\pm 1 \times 10^{-6}$ per year after 30 days operating, typical Phase Noise L(f) 100 Hz -125 dBc/Hz 1 kHz -150 dBc/Hz 10 kHz -165 dBc/Hz 20 kHz -165 dBc/Hz Temperature Stability $\pm 5 \times 10^{-7}$, 0° to +50°C (Ref +25°C) MECHANICAL Dimensions 1.5 x 1.5 x 0.5" Connectors Solder pins on base Packaging Sealed steel can POWER REQUIREMENTS Warm-Up Power 5 Watts for less than 3 minutes Total Power 2.0 Watts at +25°C Supply Voltage +12 to 15 VDC ADJUSTMENT Mechanical Tuning OR $\pm 4 \times 10^{-6}$ Electrical Tuning $\pm 5 \times 10^{-7}$, ± 5 VDC Negative slope CRYSTAL Type 100 MHz SC-cut</p>	<table border="1" style="width: 100%; border-collapse: collapse; margin-bottom: 10px;"> <thead> <tr> <th>REV</th> <th>DATE</th> <th>REVISION RECORD</th> <th>DWN</th> <th>AUTH</th> </tr> </thead> <tbody> <tr> <td>-</td> <td></td> <td></td> <td></td> <td></td> </tr> <tr> <td>A</td> <td>03-18-02</td> <td>T/CO, Total Power</td> <td>KH</td> <td>LR</td> </tr> <tr> <td> </td> <td> </td> <td> </td> <td> </td> <td> </td> </tr> <tr> <td> </td> <td> </td> <td> </td> <td> </td> <td> </td> </tr> </tbody> </table> <div style="text-align: center;">  </div> <table border="1" style="width: 100%; border-collapse: collapse; margin-top: 10px;"> <thead> <tr> <th>PIN</th> <th>FUNCTION</th> </tr> </thead> <tbody> <tr> <td>1</td> <td>Supply Voltage</td> </tr> <tr> <td>2</td> <td>Ground, Case</td> </tr> <tr> <td>3</td> <td>Ground, Case</td> </tr> <tr> <td>4</td> <td>RF Output</td> </tr> <tr> <td>5</td> <td>Electrical Tuning</td> </tr> </tbody> </table>	REV	DATE	REVISION RECORD	DWN	AUTH	-					A	03-18-02	T/CO, Total Power	KH	LR											PIN	FUNCTION	1	Supply Voltage	2	Ground, Case	3	Ground, Case	4	RF Output	5	Electrical Tuning
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Specification as shown with the crystal as advertised. The 500-16507B supplied, had electrical tuning which I suspect is 0 to -5V. On frequency is approx. -2.0V.



Noise as measured in dB/Hz with 200kHz span, note reference is +10dBm so wideband noise displayed is circa -130dBc/Hz, this is limited by analyser noise performance



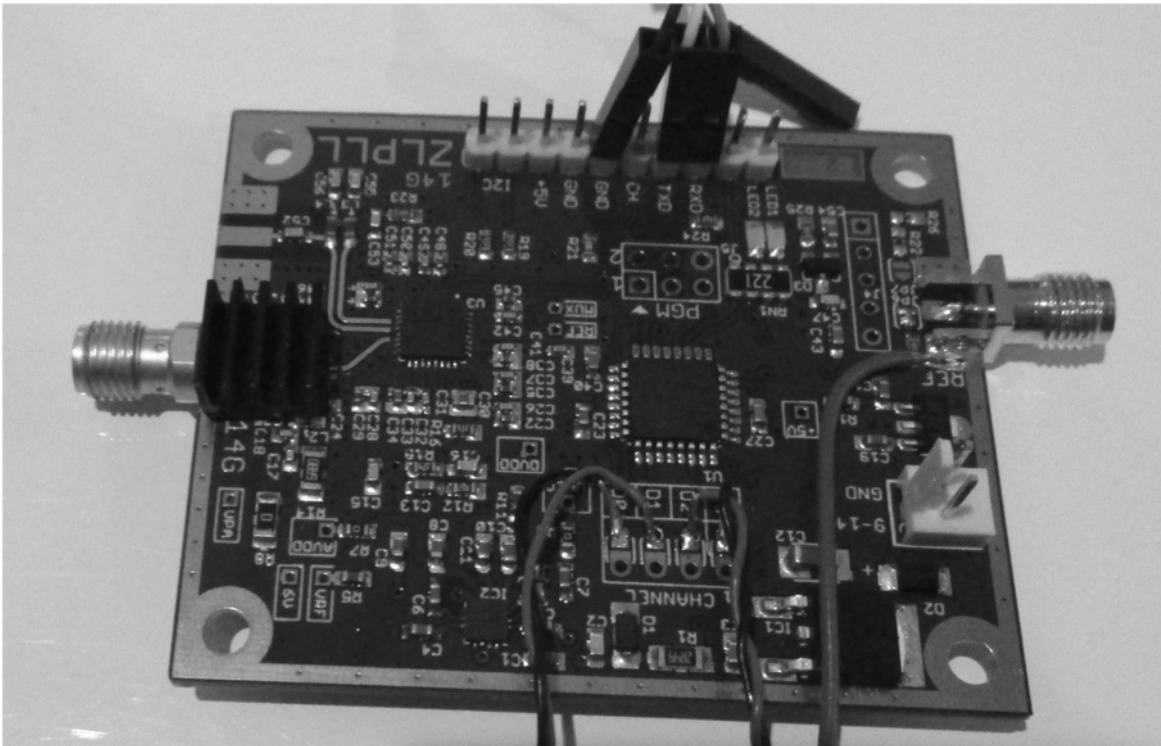
Noise in dB/Hz 200Hz span. Spurs at 50/100Hz are pickup on the bench. Ignoring those noise measured is circa -120 dBc/Hz down on the fundamental at +/- 100Hz.

Equivalent noise at 10MHz would be -140dBc/Hz.

To get stability a regulated supply is required. Specification is 12 – 15V, although measurement by John G8ACE suggest operation down to 9V. A low drop 10V regulator is the answer for portable operation from 12V. A negative voltage is required for the tune pin. This can easily be obtained in a low noise way, by using a small 3.6V lithium battery, as minimal current is required.

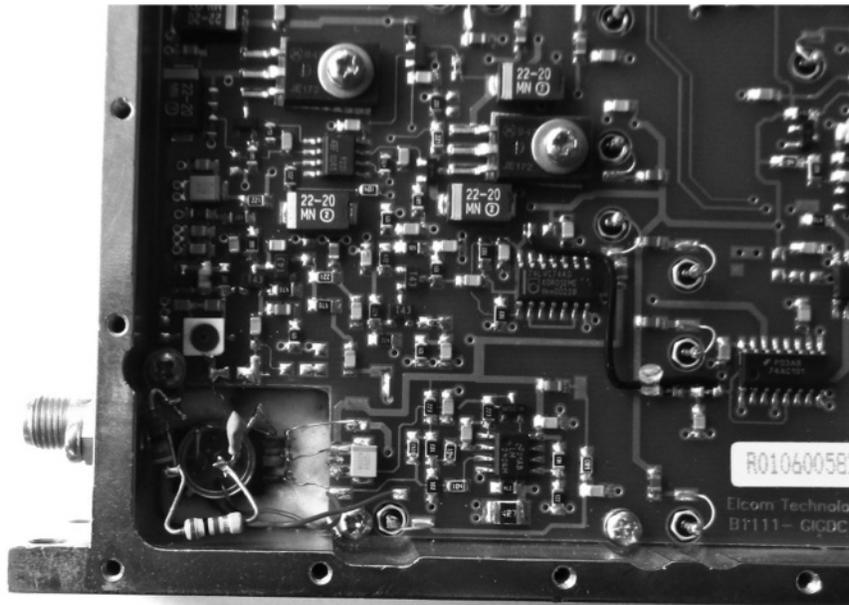
A look at the phase noise performance of a range of synthesisers and references

Part 2. Roger Ray G8CUB

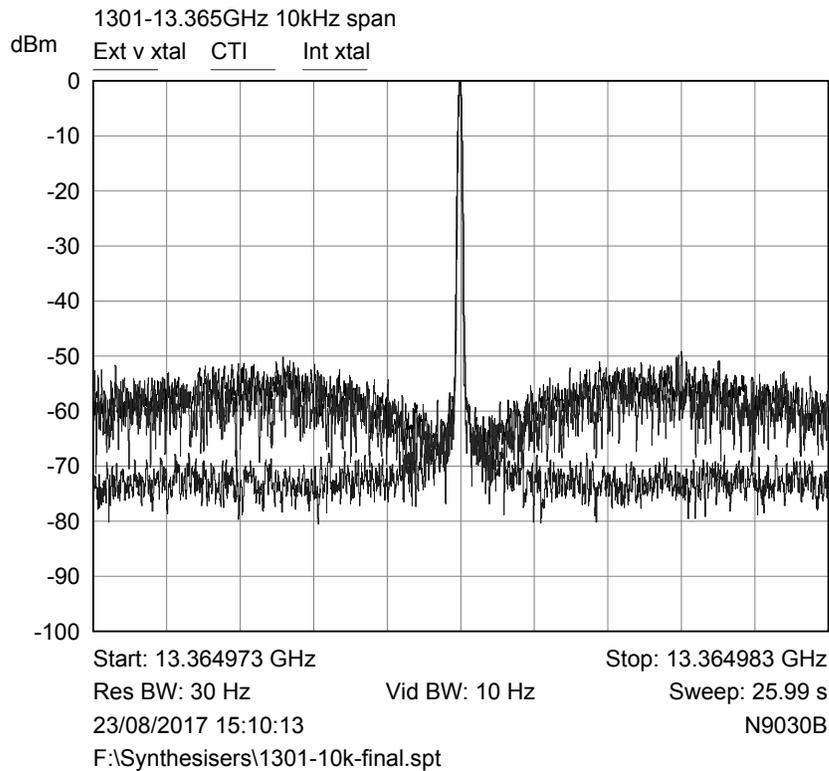


Following on from the first part of the article in November Scatterpoint. I wanted to do a few comparison plots between the different synthesisers. It has been possible to measure them all at the same frequency, or even with the same setup. So where differences in the measurement has been made, this is indicated in the caption.

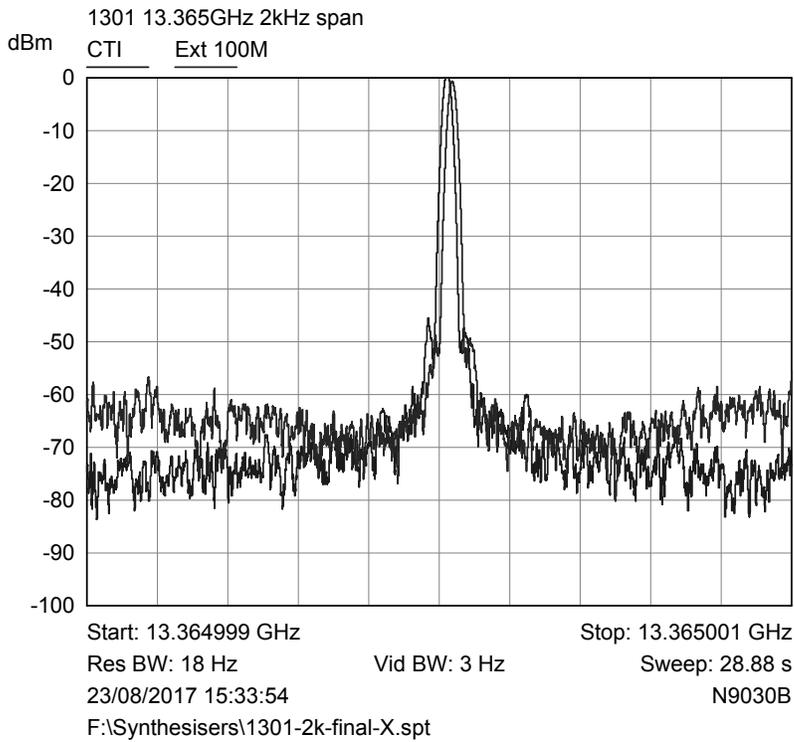
As a starting point, I looked at the Elcom DFS1301 type synthesisers. These use a 100MHz crystal with heater. Both the stability and accuracy are somewhat poorer than the later CDFSLS series that use a 10MHz TCXO.



The modification of the DFS1301, was to add an SMA connector, connected to a 47 ohm load resistor. The SMA centre pin is then connected via a 100R and 1n0 capacitor to one end of the 100MHz crystal, removed from the collector of the oscillator.

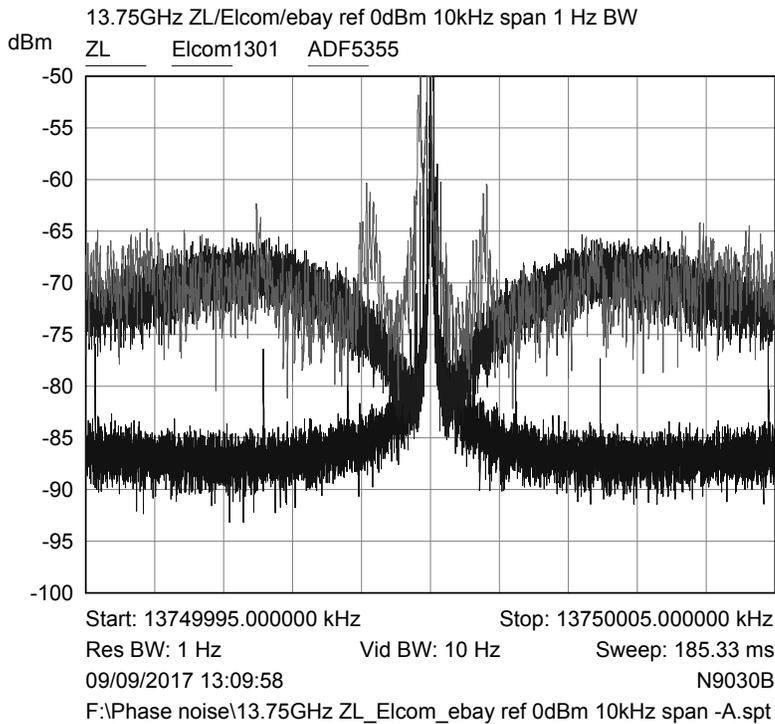


Elcom DFS1301 Internal/External 100MHz, compared to CTI Herley PLL DRO



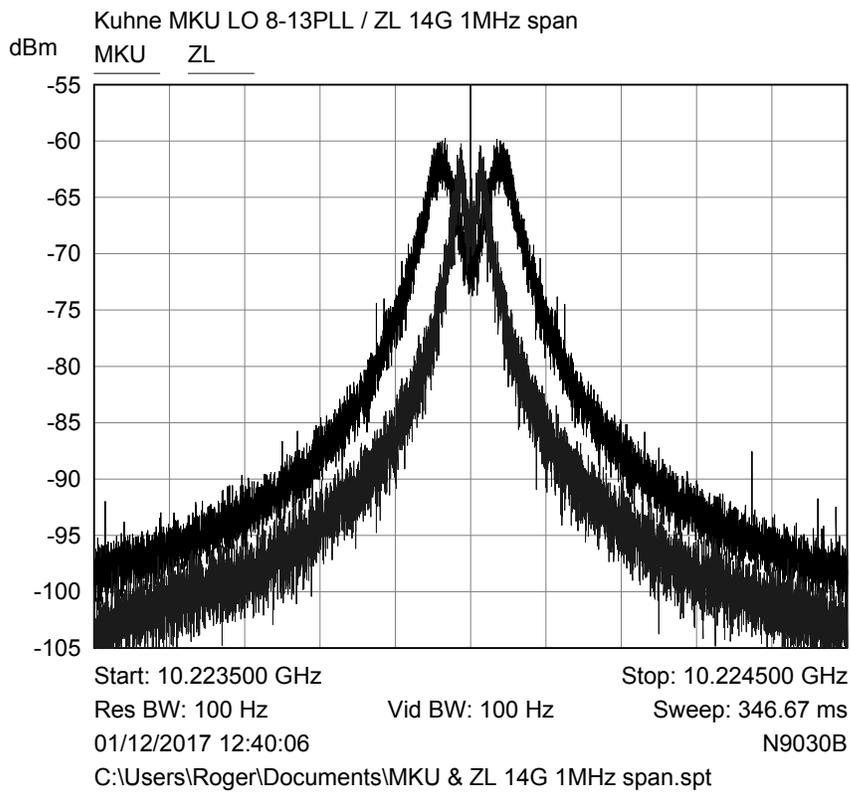
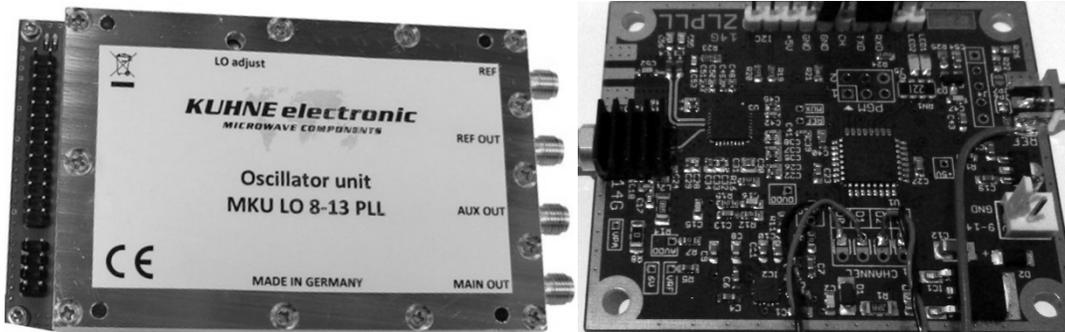
Comparison with +/- 1kHz span 18Hz BW

The blue trace is the CTI Herley @ 13.2GHz locked to a 100MHz reference. The Elcom noise was no better on the external reference (most likely limited by noise on the control line), than the internal – however it was now on frequency!

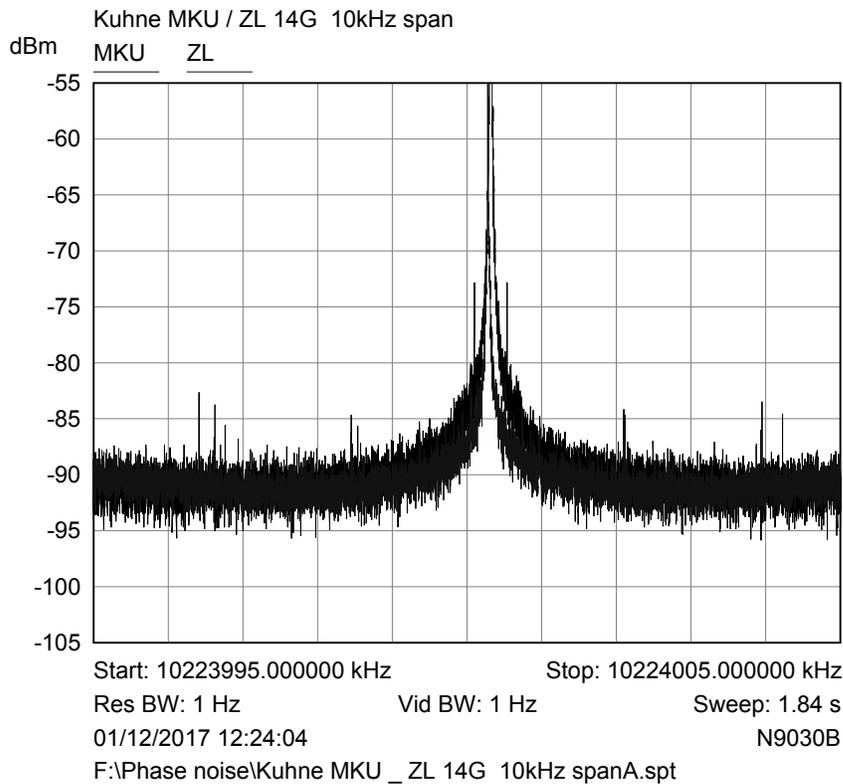


Comparison of ebay ADF5355 board with Elcom 1301 and ZL 14G synthesisers

Measurements are on the Kuhne Electronics MKU LO 8-13G PLL, and the ZL 14G synthesiser. Also included is one of the low cost ebay boards. The ebay board is shown without some of the excellent ideas of improvement by Brian Flynn and others, which mainly target the noise from the voltage regulators.



Comparison of Kuhne MKU & ZL 14G +/-500 kHz span.

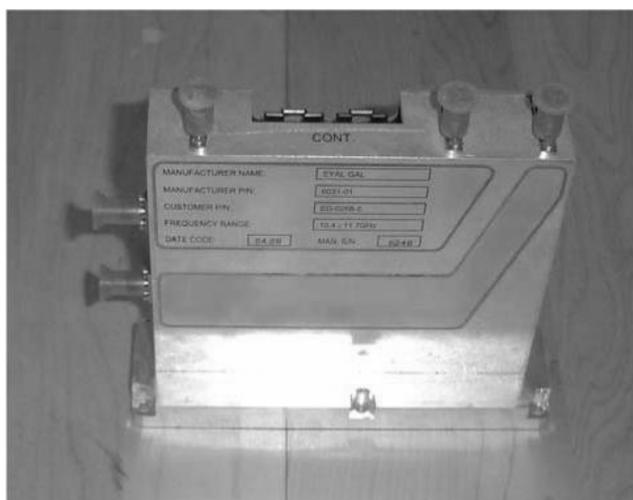


Comparison of Kuhne MKU & ZL 14G +/-10kHz span

So for a performance synthesiser, the choice is between a modified green or black ebay ADF4355, a ZL 14G, or a MKU LO. I am still waiting for the crowd funded 15GHz ERA synthesiser. Although that is in the 'reasonably priced' performance signal generator category.

My personal favourite is the ZL 14G from ZL2BKC, being reasonably priced at \$190, and including 16 programmed frequencies. Switching between frequencies can be by hex switch, or on board jumper. The programmed frequencies can easily be changed using a USB to TTL RS232 lead (a standard USB/RS232 does not work). A cheap £3.90 ebay lead worked fine. The reference input can be changed from 10 to 100MHz (or anything between), which is a bonus. There is even an on-board keyer, which allows the synthesiser to be used as a beacon.

That's not to say I don't like the Kuhne MKU, especially when used with a touch sensitive LCD controller. It is just that, even with swopping synthesisers between transverters, I need a quite a few units.



Eyal Gal 11GHz Transceiver

by Roger Ray, G8CUB

These units with the part number 6031-01 will work un-modified at 10.368GHz. They consist of a receive LNA & mixer, plus a transmit amplifier & output monitor. Thus just needing an LO, Tx mixer and filter, and a couple of relays to make a simple 10GHz transverter.

Measured performance on receive with a 432MHz IF, is as follows: Conversion gain +22dB

System noise figure 2.9dB

Image rejection (9.94GHz LO) -24.5dB

Performance on transmit:

+32dBm output - 1dB compression

+53dB gain

Saturated power output on transmit is >2W.

Supply requirements:

	Receive (Tx Inhibited)	Full Output
+8.0V	380mA	720mA
+12V	50mA	1.93A
-12V	105mA	105mA

The IF response is flat 75-1700MHz, but a 144MHz IF would only give you 2dB image rejection, so is not practical. With the high transmit gain, it is necessary to either use an input attenuator, or use the AGC control to turn the gain down.

In my 10GHz system I use the Eyal Gal block with an Alcatel synthesiser and 10MHz reference. The block dissipates a fair amount of heat, so don't remove the aluminium slab base plate.

As you can see in the picture of my complete transverter I have done just that! But, I have replaced it with an L-bracket to the base plate, which works just as well.

A specification sheet was available on the web www.eyal-microwave.com/eyal-emi/09042006100536@6031-00.pdf but no connection details, so I had to work those out.

The output power measured was much greater than that indicated by the specification sheet (harmonics are quoted with a +26dBm power, and IP3 as >+38.5dBm). I can only put that down to the fact that I am using it at the bottom end of it's frequency range.

Pin-out on the two connectors as pictured above, left to right, is as follows.

Connector 1 (6 way)

1	2	3	4	5	6
n/c	+8V	-12V	0V	n/c	+12V

Connector 2 (5 way)

7	8	9	10	11
n/c	Tx	Det	AGC	0V
	Tx	Mute		

Connections to the unit are by two, 0.1" pitch single in line connectors. It is quite easy to solder wires straight on if you don't have suitable mating plugs.

In my system, the 8.0V supply is provided by a 7808 regulator. The 12V supply is a direct battery feed. Both are internally regulated, so the exact voltage is not critical. The -12V is provided by a block DC-DC converter.

I have not had the courage to remove the -12V supply to see if the unit is internally protected, but equally I have not blown one up yet!

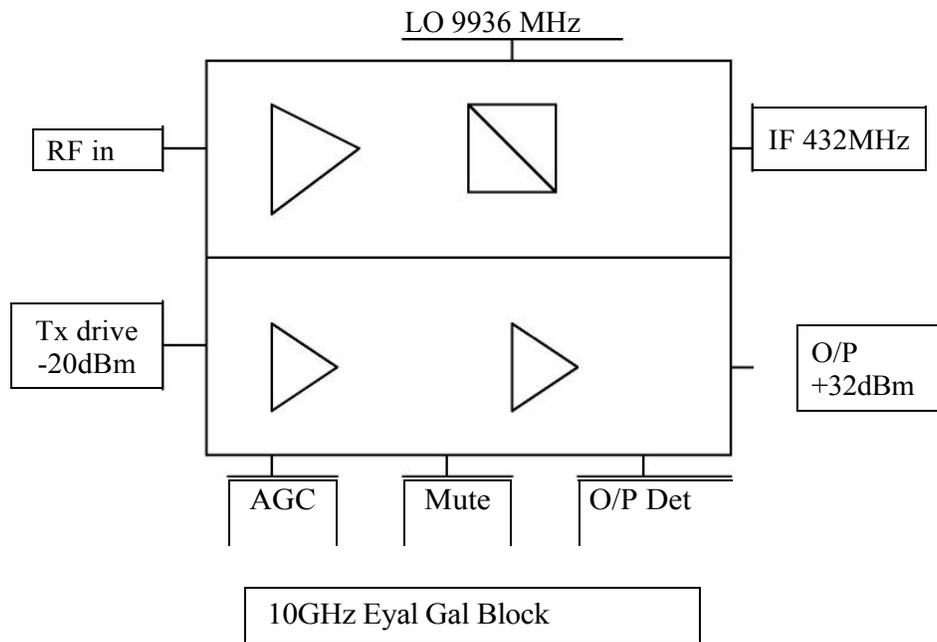
Tx Mute – 0V to inhibit

Tx Det – DC proportional to dB output power (log detector) Max ~ 4.3V

AGC – 0- 5V Control - from the data sheet turning the power down more than 10dB will limit the output power (I have not tried this, just left the pin o/c)

Local oscillator power required is +3 to +6dBm (9.936GHz). Around -21dBm Tx drive will give you full output (at max gain). The 2 SMA output connectors allow easy use of a coax relay, or the addition of a single stage low noise amp and/or PA.

Block diagram



Complete early transverter in operation

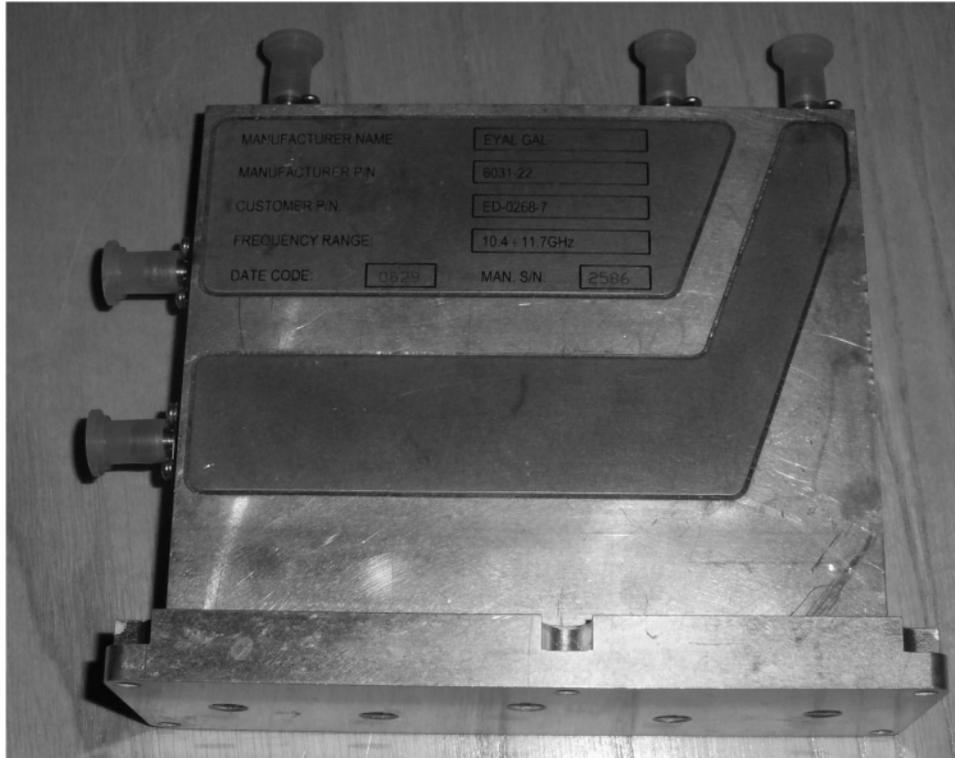


Eyal Gal Modules for 10 & 24GHz – Update

by G8CUB

My original articles on these modules in Scatterpoint is now several years old. In the intervening period, there have been periods when modules were unavailable, and others when new variants have come to light.

11GHz Module



Of the 6031 module for 10GHz, there are differences in module thickness, and derivation of transmit supply.

These 10.4 – 11.7GHz modules have a useful cut-off of the transmit response on the LF side and receive filtering is by the use of image rejection mixing / filtering.

The original modules looked at were -01 or -07 variants. These had the thicker bodies, and derived the transmit supply from the +12V.

More recent variants type -12 have the thinner bodies but still use the +12V for TX. The -22 variant has the thinner body, and takes the transmit supply from the +8V. Also the -22 types have come with test sheets showing the measured noise figure to be 1.3 to 1.6dB at 10.4GHz.

These must be first stage noise figures, as my measurements on the whole module vary between 2.3 & 3.3dB.

On the transmit side the unit is just a high gain amplifier. The 1dB compression point, measured now on quite a lot of modules, varies between 31.3 and 32.3dB, with the saturated output about 0.6dB higher. Although I have seen up to +33.5dBm.



-01 Module on the left, -22 module on the right

When using the -22 variant modules, the +8V supply needs a capability of at least 2.2A for maximum output, hence it is convenient to use a 3A regulator (78T08 or similar). This regulator needs to have sufficient heatsinking. Typical current on receive is 0.38A, while on transmit current is around 1.8A, peaking to 2.2A.

Also the -12V current requirement is less, around 70mA max. instead of 110mA. Current from the 12V supply is typically 20mA on receive, 780mA on transmit. Note that on 'transmit' i.e. TX un-muted, the receive side is still operational. The only module I have known to have a receive failure, is when the auxiliary antenna contact in the RX position was used to mute the module. Obviously the changeover period caused the problem, when using CW break-in. I have now used the TX aux. contact to control the mute line via a transistor.

Transmitter Report

Normal: PASS

Transmitter Type: 11G

CERAGON Serial No:

CERAGON Part No:

Tester Name: Alex

EY-98867
ED-0268-7

Vendor Code: 25

Vendor Serial No: 2874

Vendor Part No:

Date: 09-01 Dec. 20 2006

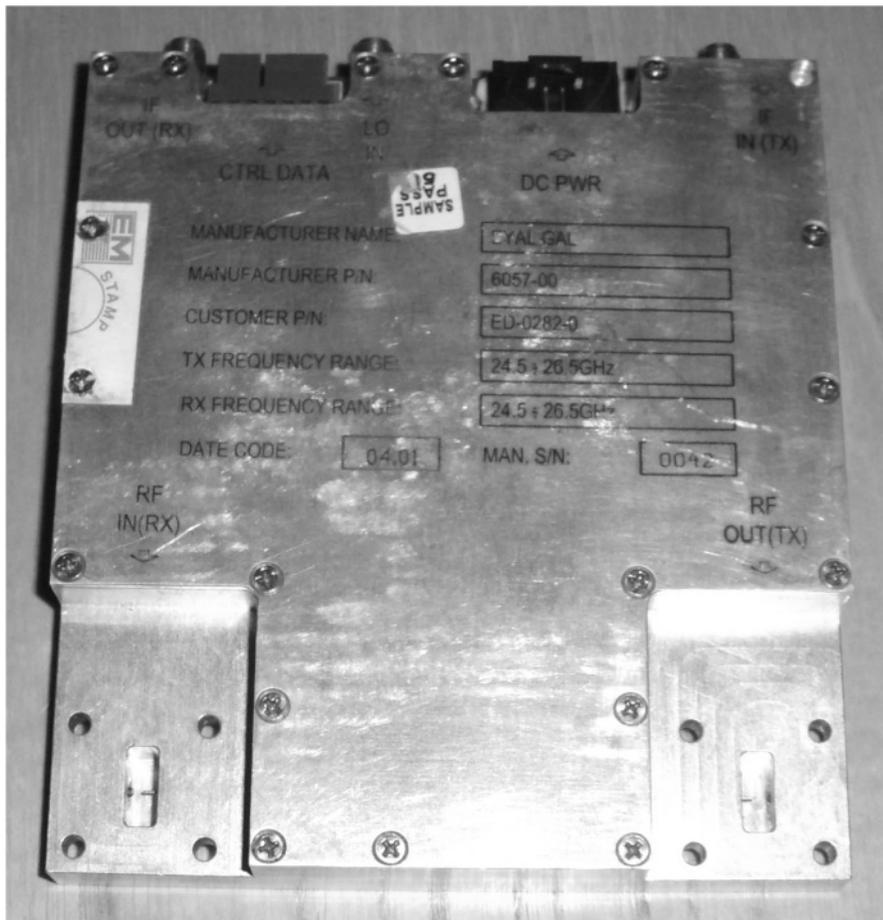
Temperature = 25°C

8031-22

DC Current: PASS	Current	Pass	Units	Pass Value
8Volt	1780	PASS	mA	
12Volt	80	PASS	mA	n < 2500 < 2500
12Volt	370	PASS	mA	n < 150 < 150
12Volt	IF	Pass	mA	n < 1800 < 1800
Tx Results: PASS			Pass	Units
Nominal Gain: PASS				Pass Value
Flatness: PASS				
IF Flatness (Max.)	0.61	PASS	dB	n < 1.2 < 1.2
Maximum Gain (Min. Val. Over Freq Range)	58.42	PASS	dB	L > 48 > 48
			dBm	
Mute: PASS				
Mute Delta	76.71	PASS	dB	n > 70 > 70
Detector: PASS				
Detector Function Derevative	90.5	PASS	mV/dB	n > 40 > 40 Above 0dBm
Detector Flatness Derevative	0.33	PASS	dB/60MHz	n < 1 < 1dB/60MHz
Gain Control: PASS				
Control Derevative	84.28	PASS	mV/dB	n > 40 > 40 Above 33dB Gain
Minimum Gain (Max. Val. Over Freq Range)	22.5	PASS	dB	n < 33 < 33
Noise Figure: PASS				
Noise Figure Start-Center	6.2		dB	
Noise Figure Center-Stop	8.5		dB	
Noise Start-Center	-71.77	PASS	dBm	n < -63 < -63
Noise Center-Stop	-69.47	PASS	dBm	n < -63 < -63
IP3/IM5: PASS				
			dBm	
			dBm	
			dBm	
Normal IP3 Start	40.3	PASS	dBm	n > 38.5 > 38.5
Normal IP3 Center	40.7	PASS	dBm	n > 38.5 > 38.5
Normal IP3 Stop	41.3	PASS	dBm	n > 38.5 > 38.5
Normal Delta IM5 Start	74.7	PASS	dB	n > 58 > 58
Normal Delta IM5 Center	71.3	PASS	dB	n > 58 > 58
Normal Delta IM5 Stop	70.6	PASS	dB	n > 58 > 58
Rx Results: PASS	IF	Pass	Pass	Units
Flatness: PASS				Pass Value
RF Flatness (Max.)		0.55	PASS	dB
Maximum Gain (Min. Val. Over Freq Range)		21.91	PASS	dB
				19.5 < Gain < 24
			mV/dB	
			dB	
Noise Figure: PASS				
Noise Figure Start		1.3	PASS	dB
Noise Figure Center		1.4	PASS	dB
Noise Figure Stop		1.4	PASS	dB

* IF Start: 10400MHz, IF Center: 11050MHz, IF Stop: 11700MHz.

Version: 5.80, Calibration Date: 17:28 Dec. 07 2006.



For 24GHz the original 6058 (21.2-23.6GHz) module I looked at, required an LO on the high side, together with additional TX filtering (when using a 70cm IF). Now there is availability of a 24.5-26.5GHz module. As with the 23GHz version these are a complete mixing system on transmit and receive, with LO at half frequency. Being designed for 26GHz allows an LO on the low side, and can be used with either no or minimal additional filtering. They do have some gain reduction at 24.05GHz. But, fortunately the waveguide connects directly into the front end amplifier via a short length of track, so the noise figure remains good. (my conservative system measurements show better than 4.5dB). They do cut-off rapidly after 24.0GHz though. After measuring a dozen units the RX gains varies between 20 and 27dB @ 24.048GHz. An advantage of this is that the image rejection on receive is >30dB.

While the unwanted product on transmit is <-40dBc, LO being around -32dBc. These measurements were made with an LO injection level of +9dBm.

They work well with the Elcom 11.2-12GHz synthesisers, just needing around 6dB of attenuation. The only disadvantage of using the Elcom, is they are limited to 3.333MHz steps. Therefore using a 70cm IF, the range is either 428-430MHz or 434.666-436.666MHz. A higher IF is practical, but not significantly lower. After using the Elcom synths at 76 & 134GHz, I recommend using the internal 10MHz reference, unless you want to go GPS locked. It is rather good once set on frequency, with quick warm-up.

Transmit measurements show a 1dB compression of around 30.2dBm. The TX gain is high, with some units only requiring -28dBm for full output. Typical supply currents are -12V 120mA, +8V 680mA RX / 1.2A TX, +12V 20mA RX / 0.65A TX.

Notes

Notes



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