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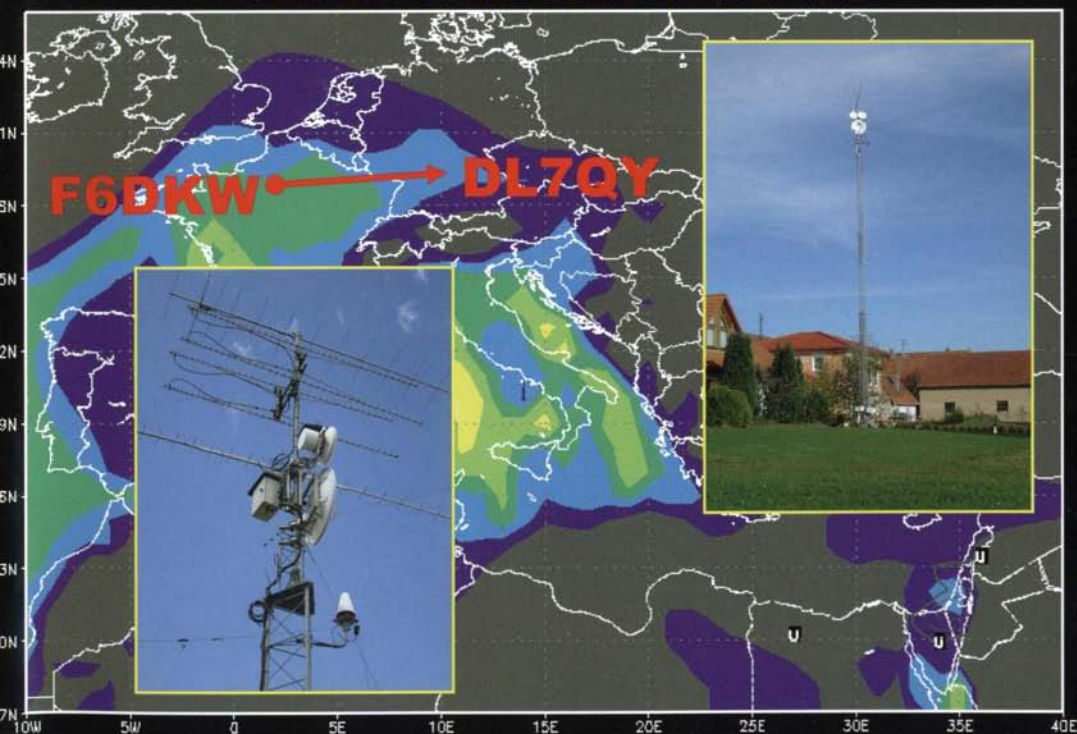
Vol.40

1. Quarter

New Tropo World Record

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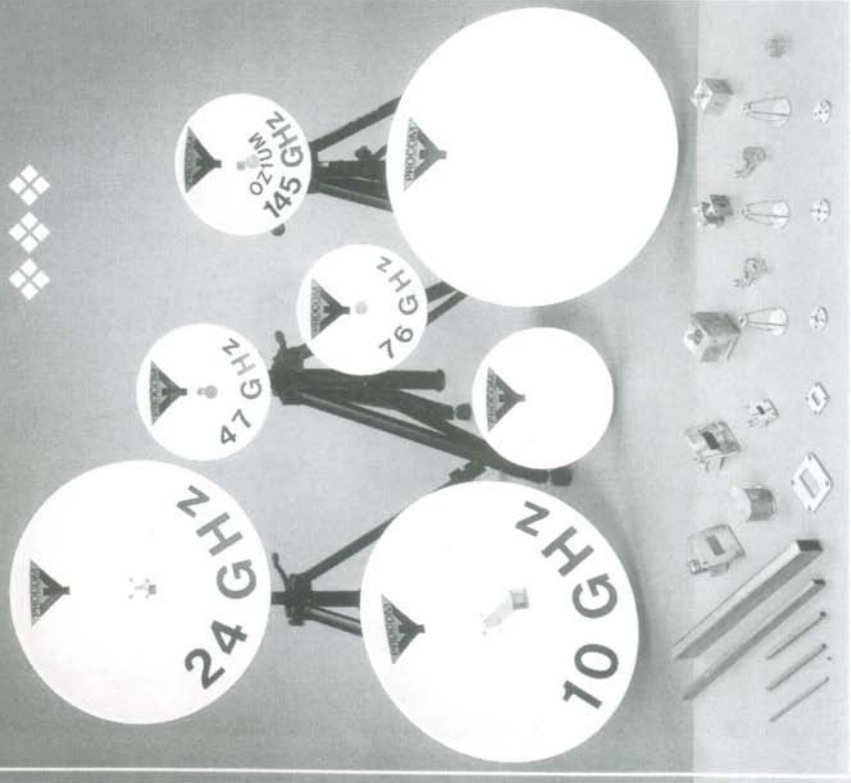
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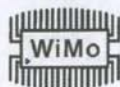
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Low Noise VHF/UHF Yagis, designed by YU7EF

High performance yagi antennas designed by Ljubisa Popa, YU7EF, optimized for a low noise figure. These antennas are well known among EME enthusiasts. Up to now, all those antennas had to be homebrewed.

But now, YU7EF has signed a contract with WiMo for production and Europe-wide sales!

We will implement the antenna designs one by one, starting with the most-wanted models.

The antennas use folded dipoles with N female socket, the elements are mounted on top of the boom. All 70cm antennas have the mast clamps below the boom to avoid influence on the radiation pattern.

The elements of the 2m yagis are made from 8mm dia. alu tubing, for the 70cm antennas we use 6mm solid aluminium rods. All support struts are V-shaped, resulting in maximal efficiency, even better than a double boom. Support struts can be mounted above of the antenna too, making the construction of yagi arrays much easier.

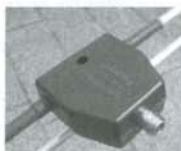
The mounting hardware for the struts are made from laser-cut, stainless steel VA sheet metal. These brackets can also be used as anchors for guying lines above the antenna, in case the struts are too heavy or impossible to use.

To compensate for the weakening of the elements due to the hole, each element is reinforced with a stainless steel shell.

The boom joints are made as high precision laser-cut sheet metal with lowest tolerances.

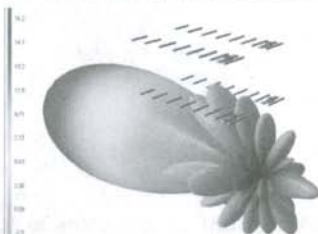
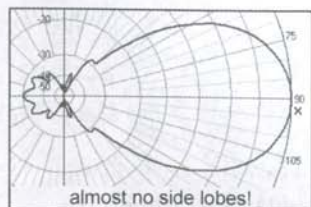
The booms are machine with very low tolerances on a CNC machine. All u-bolts, brackets, nuts and bolts are made from stainless steel, all nuts are self-locking. All other parts are made from stainless steel or aluminium. The larger antennas with strut contain no galvanized parts at all, while some smaller models have a galvanized mast bracket, otherwise are stainless/aluminum only, too.

The radiators are folded dipoles with teflon cable balun and N connector. The balun is made from 6mm teflon cable and is completely enclosed by the dipole box. The dipole box is cold welded and foam sealed for protection against humidity.



Model	Length m	Gain dBi	F/B dB	beamwidth H V		noise temp. (K)	G/T	Frequ. range (MHz)	avail- ability	price (EUR)
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EF0210LT	5,2	14,0	29	38	43	223	-9,5	143.5-145.2	autumn	199.-
EF0211B	6	15,1	29	34	37	230	-8,6	143.2-144.9	tba	239.-
EF0212	7,3	15,9	21	31	33	226	-7,7	143.8-144.6	june	268.-
EF0213M	8,0	16,0	27	39	32	225	-7,5	142.8-146	tba	289.-
EF7011N	2,2	15,3	23	33	36	28	0,8	430-435.4		148.-
EF7015	3,4	17,1	22	27	29	27	2,7	431.6-433	june	168.-
EF7015M	3,3	16,8	27	28	30	30	1,9	430-439.8		168.-
EF7027	7,3	19,5	28	21	22	23	5,8	431.5-433.1	autumn	285.-

Power rating of all antennas: 1200W (144 MHz) resp. 800W (432 MHz). Suitable mast diameter: up to 60mm.



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Editorial

Dear DUBUS Reader!

we welcome you to the first DUBUS issue of 2011!

It looks like that F1JRD's 2m PA article from last issue caused quite some interest. We got a lot of positive feedback and we guess that dozens of these PAs are currently under construction. Lionel promised to publish similar designs for 6m and 70cm in DUBUS soon. We were asked by some readers to publish a also "complete" version of the 2m PA, incl. protection circuits etc. So if anyone has completed and tested such an assembly successfully, DUBUS would be interested to publish it. F1JRD has put a video of his 2m PA under test on Youtube, that got >2000 hits within 48 hours, not bad.... Search for "freescall" and "dubus" on Youtube and you will find it.

We want to remind everybody to take part in our CW EME Contests in March, April, May, and June. The 2011 contest is dedicated to Claude Descat, F3VS, who became a silent key in January 2011. See full rules in this DUBUS issue on page 118 or on www.dubus.org. Please note that there are some changes regarding QSO points and multiband scoring!

The latest forecast for the maximum of the new solar cycle has been corrected further down. NASA experts now say that the maximum will occur even later in June/July 2013 with a mean sun spot number (SSN) of only 59 (!). This would mean that there hardly will be F2 openings on 6m on the mid latitudes on east to west paths, like central EU to NA. The latest numbers from January 2011 confirm this poor forecast, unfortunately.

Last not least we want to mention that a new book DUBUS TECHNIK X is available now via your local distributor or directly from us in Germany. The book contains the articles from DUBUS magazine 2009 & 2010 including F1JRD's PA and HB9TXV's VNA. A PDF with the complete content is on www.DUBUS.org.

Please keep sending your reports and technical articles and if you have not yet renewed for 2011, may be to do it now.

Thank you!

73 from Joe, DL8HCZ / CT1HZE
and the DUBUS team!

Liebe DUBUS Leser!

Willkommen zur ersten DUBUS-Ausgabe des Jahres 2011!

Es sieht ganz so aus, dass der 2m-PA-Artikel von F1JRD, der in der letzten Ausgabe veröffentlicht wurde, einiges Interesse hervorgerufen hat. Wir haben viele positive Rückmeldungen erhalten und glauben, dass momentan dutzende dieser PAs im Bau sind. Lionel hat versprochen, dass er ähnliche Versionen für 6m und 70cm bald in der DUBUS veröffentlichen wird. Wir wurden auch daraufhin angesprochen, eine „Komplettversion“ der 2m-PA zu veröffentlichen, bei der Schutzschaltungen usw. enthalten sind. Wenn also einer der Nachbauer so etwas fertig gestellt und getestet hat, wären wir ggf. an einem Abdruck interessiert. F1JRD hat ein Video seiner 2m-PA bei Youtube eingestellt. Einfach dort nach „freescall“ und „dubus“ suchen, dann findet man es sofort. Das Video wurde innerhalb von nur 2 Tagen über 2000 x angeschaut, nicht schlecht, oder?

Wir möchten nochmals alle erinnern, am CW EME-Contest im März, April, Mai und Juni teilzunehmen. Die Regeln stehen in dieser Ausgabe auf Seite 118 und unter www.dubus.org. Achtung: Es gibt Änderungen bei den Punkten und bei der Multiband-kalkulation!

Die neuesten Vorhersagen für das nächste Sonnenfleckmaximum wurden gerade wieder noch weiter nach unten korrigiert. Die NASA-Experten gehen nun von einem noch späteren Maximum im Juni/Juli 2013 aus, wobei von einer gemittelten Sonnenfleckenzahl von nur noch 59 (!) ausgegangen wird. Das würde bedeuten, dass kaum mit F2-Öffnungen auf 6m in den mittleren Breiten in Ost-West-Richtungen zu rechnen ist, z.B. also von Mittel-Europa nach Nordamerika. Die aktuellen Zahlen vom Januar 2011 bestätigen leider diese neue schlechte Vorhersage.

Zum Schluß noch der Hinweis, dass es ein neues Buch DUBUS TECHNIK 10 gibt. Das Inhaltsverzeichnis ist auf www.DUBUS.org zu finden. - Für die Einsendung technischer Artikel – auch kleinerer! – sind wir immer dankbar.

73 von Joachim, DL8HCZ / CT1HZE
und vom ganzen DUBUS-Team!

A home-made 76 GHz dual-mode feedhorn

by Jean-François Lampin, F4BAY - © 2011

Introduction

Dual-mode feedhorns provide excellent performance for the illumination of parabolic dishes or dielectric lenses. As the frequency increases it is more difficult for the amateur to fabricate such horns with the required precision. Recently W1GHZ has published an article [1] about optimized dual-mode feedhorns inspired by the work of Skobelev et al. [2]. The geometry of these horns is quite simple: a circular waveguide followed by a step and a flared horn. It is possible to fabricate such a horn for the lowest microwave amateur band and also for the mm bands using amateur tools. Here it is proposed to describe fabricating such a dual-mode feedhorn for the 76 GHz (4 mm) band.

Design

The goal is to illuminate a $f/D = 1$ dielectric lens. The horn is chosen from table 1 in [1]. The best design is a horn with an aperture diameter of 2.3 wavelengths and a length 10% shorter than the nominal length given by Skobelev (3.26 wavelengths). For this design, W1GHZ calculated an optimum f/D of 0.98 and an efficiency close to 80 %. The phase center is theoretically in the plane of the aperture. A sketch of the horn with the theoretical critical dimensions is shown in figure 1. The diameter of the input circular waveguide is larger than for standard single-mode waveguide. Instead a circular input waveguide made of brass tubing with an internal diameter of 3.0 mm was used. A flared section to the 4.0 mm waveguide is then needed.

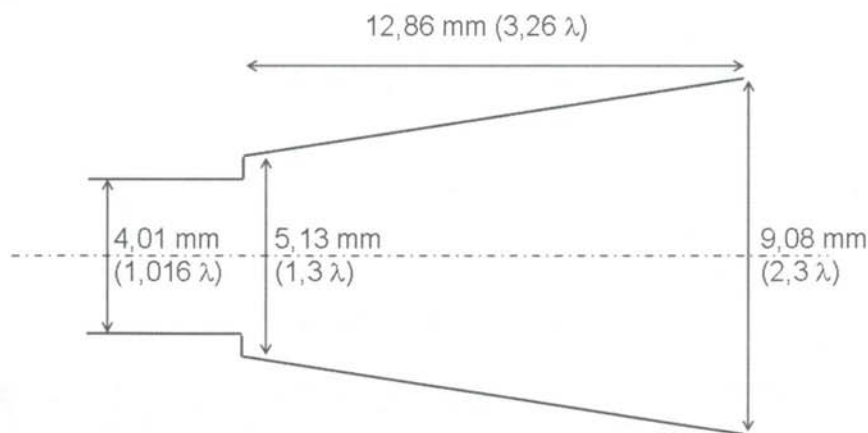


Fig. 1: Sketch of the 76 GHz dual-mode feedhorn

Mechanical construction

The mechanical drawing of the horn is shown in figure 2. It is composed of five elements: an input waveguide, a flange, a cylindrical section, a flared section and a ring. The input waveguide is made of standard brass tube with an internal diameter of 3.0 mm and an external diameter of 4.0 mm. A 22 mm length is cut and its end flared with a semi-angle of about 20° to obtain a final diameter of 4 mm using a "high speed cutter" (model 125 from Dremel). The flange is made from brass stock on a lathe and drilled at the center for the 4 mm tube (figure 3). Four holes are drilled and threaded for the screws (M2 or 4-40 UNC). Two other holes are drilled for index pins (if needed). The cylindrical section is made from a 8mm long, 6 mm diameter brass rod drilled at the center with a diameter of 4.0 mm.

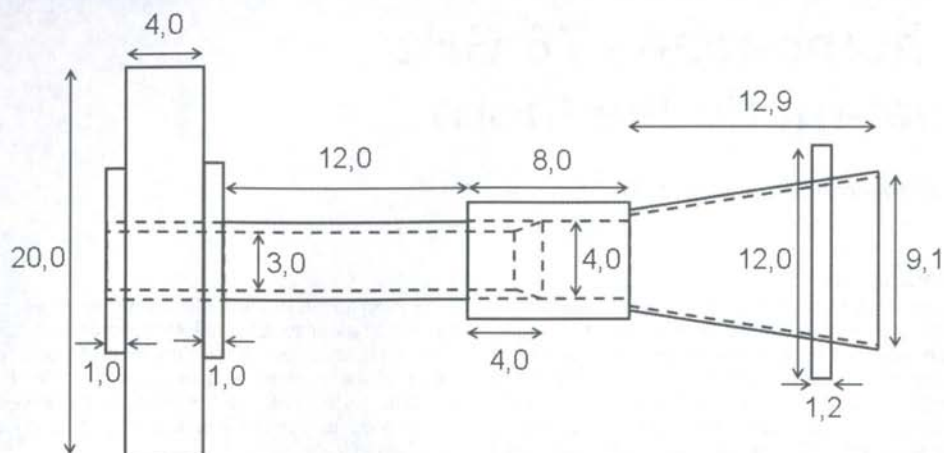


Fig. 2: Dimensions of the home-made dual-mode horn

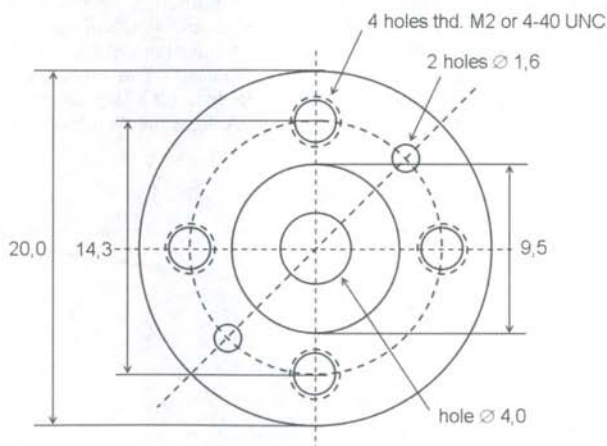


Fig. 3: Dimensions of the flange

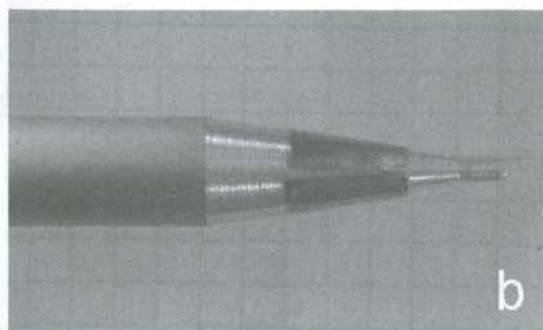
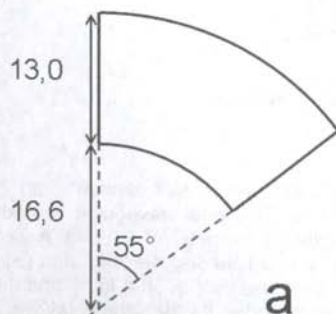


Fig. 4: Dimensions of the copper sheet (a), view of the sheet rolled around the aluminium cone (b)

The flared section is made from a shape precisely cut from 0.25 mm-thick copper sheet (figure 4a). This shape has to be rolled around a cone to precisely form the horn. This cone of semi-angle 9° was fabricated on a lathe from a 12 mm diameter aluminium rod. The horn is then formed (figure 4b) but it is generally difficult to completely close the horn due to the elasticity of the metal. A ring is used to help closing the horn during the assembly. It is made from a 1.2 mm thick brass washer (external diameter: 12 mm, internal diameter: 9 mm).

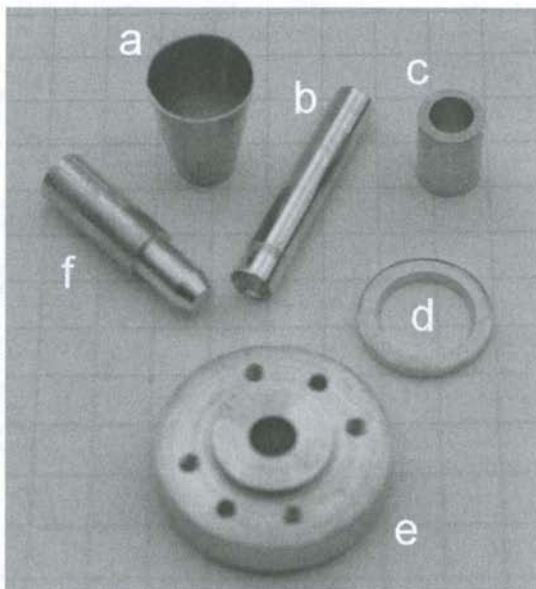


Fig. 5: Elements of the horn before soldering. Flared section (a), input waveguide (b), cylindrical section (c), ring (d), flange (e), centering tool (f)

When all the elements are ready and cleaned (figure 5a-e), the horn is assembled. First, the input waveguide is inserted into the flange and soldered using a standard lead-tin alloy solder. Then the other end of the input waveguide is inserted into the cylindrical section and soldered. Now the ring is inserted around the copper flared section in order to close it. The ring and the two edges of the flared section are soldered. It is preferable to use only the minimum quantity of solder, avoided letting the solder penetrate inside the horn. Finally the horn is soldered on the cylindrical section. In order to center the horn precisely on the cylindrical section, a centering tool is made from a 5 mm diameter aluminium rod (figure 5f). A step with a diameter of 4 mm is formed using a lathe. The tool is fitted during the soldering and then removed with the help of tweezers. Figure 6 shows different views of the fabricated horn.

Measurements

Characteristics of the dual-mode feedhorn were measured using 76 GHz waveguide equipment. First, the return loss was measured with a 10 dB-WR15 directional coupler. A WR15 to WR12 taper transition was used but the flange of the horn was directly fitted to the WR12 flange (no circular/rectangular transition). A return loss of about 12 dB was measured in this configuration. This value is reasonable and would be probably lower if a rectangular/circular transition was used. Then, the radiation patterns were measured using a X-band rotary joint, a 76 GHz $\times 8$ multiplier (DB6NT) and the horn under test for the emitter side; a pyramidal horn and a zero bias detector diode for the receiver side. It was carefully checked that the detector diode was working in its square-law region. The H-plane pattern was measured by fitting the horn directly on the multiplier. For the E-plane, a waveguide twist was added and the detector was rotated by 90° . Figure 7 shows the measured radiation patterns with a dynamic range of 30 dB. The main lobe is very clean and symmetric. With this dynamic no secondary lobes were detected which is coherent with the calculations of W1GHZ [1]. The measured beamwidth at -10 dB are 56° (H-plane) and 50° (E-plane). These values are very close to the value calculated by W1GHZ (56°). The small difference between the two planes is probably due to small mechanical imperfections and is quite acceptable.

Summary

A 76 GHz dual-mode feedhorn was successfully fabricated using only standard tools. Solder is used for assembly and no high-cost precision machining is needed. A lathe is required only for the aluminium cone and the centering tool which can be easily fabricated and re-used for the fabrication of several horns. The flange can be purchased or fabricated with a drill if a lathe is not available. This horn gives good illumination of apertures whose f/D in the 0.8-1.2 range, with an optimum around 1.0. If other f/D values are needed, the feedhorn can be easily modified by changing only the output flared section (see [1]). The cone and the shape of the copper foil must be modified accordingly.

References

- [1] Paul Wade, W1GHZ, Optimized dual-mode feedhorns, Dubus 3/2009, Vol. 38, pp. 46-65 & "Technik 9"
- [2] S.P. Skobelev, B.-J. Ku, A.-V. Shishlov, and D.-S. Ahn, "Optimum geometry and performance of a dual-mode horn modification", *IEEE Antennas and Propagation Magazine*, Vol. 43, pp. 90-93, February 2001.

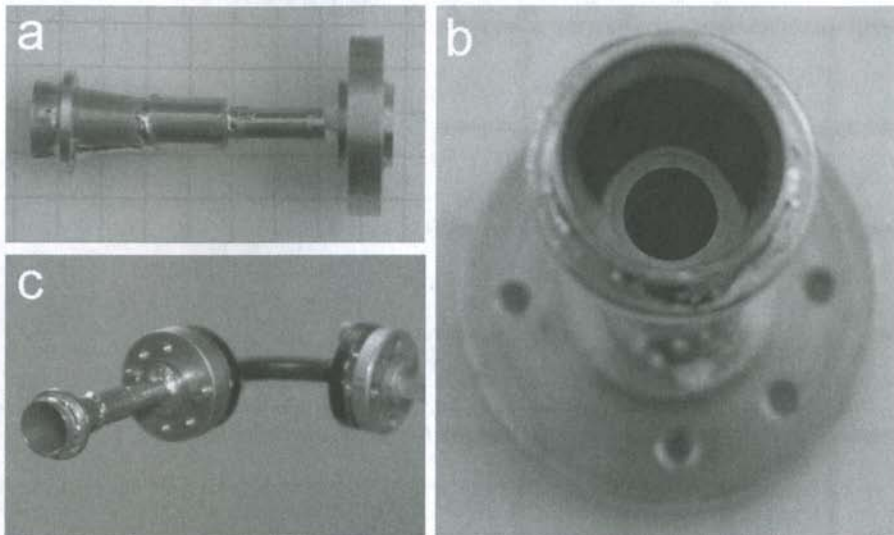


Fig. 6: The horn after soldering: side view (a), inside view (b), under measurement (c)

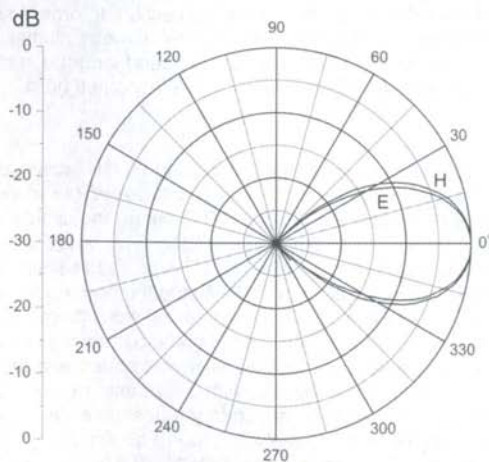


Fig. 7: Measured radiation patterns at 76,0 GHz in H-plane and in E-plane

Selbstbau eines Dualmode-Feedhorn für 76 GHz

von Jean-François Lampin, F4BAY © 2011

Einleitung

Dualmode Feedhörner eignen sich hervorragend für die Ausleuchtung eines Parabolspiegels oder von dielektrischen Linsen. Mit steigender Frequenz wird es schwieriger für den Amateur solche Hörner mit der nötigen Genauigkeit herzustellen. Vor kurzem hat W1GHZ einen Artikel über Dualmode-Feedhörner publiziert [1], angeregt durch die Arbeiten von Skobelev et al. [2]. Die Geometrie dieser Hörner ist recht einfach: einem zirkularen Wellenleiter folgt eine Stufe und ein aufgeweitetes Horn. Es ist möglich solch ein Horn für die unteren Mikrowellen-Amateurbänder und auch für die Millimeter-Bänder mit Amateurmitteln herzustellen. Dieser Artikel beschäftigt sich mit der Anfertigung eines solchen Dualmode-Feedhorns für das 76-GHz-Band (4mm).

Design

Das Ziel ist eine dielektrische Linse mit einem $f/D = 1$ auszuleuchten. Das Horn wurde aus Tabelle 1 in [1] ausgewählt. Das beste Design ist ein Horn mit einem Öffnungsdurchmesser von $2,3 \lambda$ und einer Länge, die 10% kürzer ist als die nominale Länge, die von Skobelev ($3,26 \lambda$) vorgegeben ist. Für dieses Design hat W1GHZ ein optimales f/d von 0.98 errechnet und eine Effektivität von nahe bei 80 %. Das Phasenzentrum liegt theoretisch in der Ebene der Öffnung. Eine Übersichtszeichnung des Horns mit den theoretisch kritischen Abmessungen zeigt Abb. 1. Der Durchmesser des zirkularen Eingangswellenleiters ist größer als beim Standard-Einfachmode Wellenleiter. Ein runder Eingangswellenleiter aus Messingrohr mit einem inneren Durchmesser von 3.0 mm wurde hierfür gewählt. Eine Aufweitungs-Sektion auf den 4.0 mm-Wellenleiter wird dann noch benötigt.

Mechanische Konstruktion

Die mechanischen Abmessungen des Horns werden in Abb. 2 gezeigt. Es besteht aus 5 Elementen: Einem Eingangswellenleiter, einem Flansch, einem zylindrischen Abschnitt, einer aufgeweiteten Sektion und einem Ring. Der Eingangswellenleiter wird aus Standard-Messingrohr mit einem Innendurchmesser von 3.0 mm und einem Außendurchmesser von 4.0 mm hergestellt. Es wird ein Stück mit 22 mm Länge abgeschnitten, dessen Ende mit einem Halbwinkel von etwa 20° aufgeweitet wird, um den Enddurchmesser von 4.0 mm zu erreichen. Dies wurde mit einem „Hochgeschwindigkeitsschleifer“ (Model 125 von Dremel) durchgeführt. Der Flansch wird aus einem Messingblock auf einer Drehbank hergestellt und im Zentrum für das 4-mm-Rohr durchbohrt (Abb. 3). Vier Löcher werden gebohrt und Gewinde für die Schrauben geschnitten (M2 oder 4-40 UNC). Zwei weitere Löcher werden für die Index-Pins gebohrt, sofern nötig. Die zylindrische Sektion wird aus einem 8 mm langen Messingstab mit 6 mm Durchmesser angefertigt, in den in die Mitte mit 4.0 mm hineingebohrt wird.

Die aufgeweitete Sektion wird aus einem präzise zugeschnittenen Kupferblech aus 0.25 mm Stärke hergestellt (Abb. 4a). Dieses Blech muß um einen Konus herum gerollt werden, um die präzise Hornform zu ergeben. Der Konus mit einem Halbwinkel von 9° wurde auf einer Dehbank aus einem 12-mm-Alu-Stab angefertigt. Das Horn wird dann geformt (Abb. 4b). Aufgrund der Elastizität des Metalls ist es generell schwierig das Horn komplett zu schließen. Um das Horn während des Zusammenbaus zu schließen wird deshalb ein Hilfsring benutzt. Es wird dafür eine 1.2 mm dicke Messingscheibe (Außendurchmesser 12 mm, Innendurchmesser 9 mm) verwendet).

Wenn alle Teile fertig und gesäubert sind (Abb. 5a bis 5e), wird das Horn zusammengesetzt. Zuerst wird der Eingangswellenleiter in den Flansch gesetzt und mit Standard-Lot eingelötet. Dann wird das andere Ende des Eingangswellenleiters in die zylindrische Sektion eingeführt und verlötet. Nun wird der Ring um das noch offene Horn gelegt, um es zu schließen. Der Ring und die zwei Ränder der aufgeweiteten Sektion werden verlötet. Vorzugsweise wird hier nur die geringstmögliche Menge Lot verwendet, weil verhindert werden muss, dass zuviel Lot ins Innere des Horns gelangt. Zum Schluß wird das Horn auf die zylindrische Sektion aufgelötet. Um das Horn genau zentriert auf diese Sektion zu bekommen, wird ein

Zentrierungshilfsmittel aus einem 5mm Aluminiumstab (Abb. 5f) hergestellt. Eine Stufe mit 4mm Durchmesser wird mit einer Drehbank aufgedreht. Der Hilfsstab wird während des Lötvorgangs eingesetzt und danach mit einer Pinzette wieder entfernt. Abb. 6 zeigt verschiedene Ansichten des fertigen Horns.

Messungen

Das Dualmode-Feedhorn wurde mit 76-GHz-Wellenleiter-Equipment gemessen. Der Returnloss wurde mit einem 10dB-WR15-Richtkoppler gemessen. Ein WR15 auf WR12 Übergang wurde verwendet, aber der Flansch des Horns war direkt auf den WR12-Flansch geschraubt (kein Übergang von zirkular auf rechteckig). In dieser Anordnung wurde ein Returnloss von 12 dB gemessen. Dieser Wert entspricht den Erwartungen und würde wahrscheinlich besser sein, wenn ein Übergang von zirkular auf rechteckig verwendet würde. Dann wurden die Strahlungsdiagramme mit einer Drehverbindung fürs X-Band, einem Verachtfacher für 76 GHz (DB6NT) und dem zu testenden Horn auf der Sendeseite, ausgemessen. Ein pyramidales Horn und eine Zerobias-Detektor-Diode wurden beim Empfänger eingesetzt. Es wurde sorgfältig sichergestellt, dass die Detektor-Diode nur im quadratischen Bereich arbeitete. Das Diagramm der H-Ebene wurde mit dem direkt auf den Multiplier geflanschten Horn gemessen. Für die E-Ebene wurde eine Drehung eingefügt und der Detektor um 90 Grad gedreht. Abb. 7 zeigt die gemessenen Strahlungsdiagramme mit einem Dynamikbereich von 30 dB. Die Hauptkeule ist sehr sauber und symmetrisch. Bei diesem Dynamikumfang wurden keine Nebenzipfel beobachtet, was mit den Berechnungen von W1GHZ [1] übereinstimmt. Der gemessene Öffnungswinkel für -10dB beträgt 56° (H-Ebene) und 50° (E-Ebene). Diese Werte liegen sehr nahe bei den von W1GHZ errechneten Werten (56°). Der kleine Unterschied zwischen den beiden Ebenen rührt wahrscheinlich von kleinen mechanischen Ungenauigkeiten und ist recht akzeptabel.

Zusammenfassung

Ein Dualmode-Feedhorn für 76 GHz wurde erfolgreich nur mit Standard-Werkzeugen hergestellt. Es wurde gelötet und keine teuren Präzisionsmaschinen benötigt. Eine Drehbank ist nur nötig, um den Aluminium-Konus und den Zentrierungsstab zu drehen. Dies ist einfach und beide können wieder verwendet werden, um weitere Hörner herzustellen. Der Flansch kann gekauft werden oder nur mit einem Bohrer hergestellt werden, wenn keine Drehbank zur Verfügung steht. Dieses Horn hat sehr gute Eigenschaften für die Ausleuchtung in einem f/D-Bereich von 0.8 bis 1.2, bei einem Optimum um 1 herum. Wenn andere f/D-Werte benötigt werden, kann das Feedhorn leicht modifiziert werden, indem lediglich die aufgeweitete Output-Sektion abgeändert wird (siehe [1]). Der Konus und das Kupferblech müssen dann entsprechend geändert werden.

Referenzen

- [1] Paul Wade, W1GHZ, Optimierte Dualmode-Feedhörner, DUBUS 3/2009, Jg. 38, S. 46-69 und DUBUS TECHNIK IX, S. 74-97
- [2] S.P. Skobelev, B.-J. Ku, A.-V. Shishlov, and D.-S. Ahn, "Optimum geometry and performance of a dual-mode horn modification", *IEEE Antennas and Propagation Magazine*, Vol. 43, pp. 90-93, February 2001.

Abbildungen

Abb. 1 – Übersichtszeichnung des 76-GHz-Dualmode Feedhorns

Abb. 2 – Dimensionen des selbstgebauten Dualmode-Horns

Abb. 3 – Dimensionen des Flansches

Abb. 4 – Dimensionen des Kupferblechs (a),

Ansicht des Blechs, das um den Alu-Konus gelegt ist (b)

Abb. 5 – Elemente des Horns vor dem Zusammenlöten.

Aufgeweitete Sektion (a), Eingangswellenleiter (b),

zylindrische Sektion (c), (Hilfs-)Ring (d), Flansch (e), Zentrierungshilfsstab (f)

Abb. 6 – Das Horn nach dem Zusammenlöten:

Von der Seite (a), innen (b), bei der Messung (c)

Abb. 7 – Gemessene Strahlungsdiagramme auf 76,0 GHz in der H- und E-Ebene

Hybrid-LFA-Oblong-Yagis

von Slobodan Bukvic, YU7XL

Einleitung

Gibt es jemanden, der IK3MAC noch nicht via Mond gehört hat? Naja, der Grund ist seine 24 x 23 Element Monster-Antenne für 144 MHz, die den Job erledigt. Es gibt einige Fotos dieser Anlage im Internet. Man erkennt, dass es sich um eine Kombination aus Yagi- und Loop-Elementen handelt, aber es gibt keine Details dazu. Ich habe versucht das Geheimnis zu lüften.

Mein erster Schritt bestand darin, EZNEC-Files einiger guter Yagi-Designs zu nehmen und diese zu einem Hybrid umzuwandeln und zu schauen, was passiert. Ich habe einige meiner 200-Ohm-LFA-Yagis und auch einige zur LFA abgewandelten YU7EF-Yagis genommen. Zuerst, wenn nur der letzte Direktor umgewandelt wurde, wurde ein höherer Gewinn und besseres G/T erreicht. Die Fußpunktimpedanz blieb bei 200 Ohm. Jeder weitere umgewandelte Direktor erbrachte immer bessere Ergebnisse. Nicht ganz... Sobald diese Umwandlung an die dem Strahler nächsten Elemente anlangte, vergrößerte sich die Antennentemperatur wieder, und, obwohl der Gewinn ein wenig höher war, sank das G/T ab. Gleichzeitig wurde die Impedanz inakzeptabel, so dass einige Änderungen bei der Elementposition nötig waren, was das G/T sogar noch schlechter machte. Die beste Lösung, die ich fand, bestand darin, Reflektor, Strahler und die ersten 2 bis 5 Direktoren als Yagi-Elemente auszuführen und den Rest als Loop-Elemente. Und das ist genau das, was wir mehr oder weniger beim Design von IK3MAC sehen. Diese Antenne übertrifft alle existierenden normalen Yagi- oder Loop-Antennen!

Ich weiss nicht, ob IK3MAC sein Design offenlegen wird. Ich habe jedenfalls herausgefunden, dass Rechtecke mit einem Weiten-Höhen-Verhältnis von 1 die besten Ergebnisse brachten. Dieses Verhältnis ist gut für einen weiteren Vorteil, wie wir gleich sehen werden. Wahrscheinlich ist die runde Schleife (Loop) noch besser, aber für mich ist es praktisch unmöglich eine rundes Modell mit EZNAC zu machen. Ich muss betonen, dass ich diverse Antennen für 50-Ohm-Fußpunktimpedanz modelliert habe. Die Ergebnisse waren nicht so gut wie bei einem 200-Ohm-System. Deshalb haben alle meine Modelle – sofern nicht anders angegeben – 200 Ohm Impedanz und der Speisepunkt ist beim DE2. Das bedeutet, dass ein $\lambda/2$ -Koaxialkabel zur Anpassung und Symmetrierung nötig ist. Hier sind zwei meiner besten Hybrid-Modelle:

Antenne QY21105x14

Diese Antenne besteht aus 7 Yagi- und 4 Oblong-Elementen. Alles aus Alu mit 5mm Durchmesser.

Abb. 1: QY21105x14 Anordnung

Leistungsdaten & Vergleichstabelle

Tabelle 1 & 2 – Die erste Reihe gibt jeweils die Freiraumdaten ohne Verluste wieder, die untere Reihe die Daten mit den Verlusten für Aluminium.

Abmessungen

Tabelle 3 - Alle Elemente aus Alu mit $\varnothing$ 5 mm.

Die Höhe der oblongen Elemente beträgt 500mm; Daten ohne Boom-Korrektur.

Man sieht, dass R und die ersten Direktoren Yagi-Elemente sind. DQ1 bis DQ4 sind oblonge Elemente mit 500mm Höhe, die horizontale Länge ist in der Tabelle genannt. Alles Alu mit 5mm Durchmesser.

Diagramme (inkl. Verluste für Aluminium, Simulation)

Abb. 3: Azimuth plot QY21105x14

Abb. 4: Elevation plot QY21105x14

Abb. 5: SWR plots QY21105x14 (Simulation)

Antenna QY21308XL4

Diese Antenne besteht aus 7 Yagi- und 6 Oblong-Elementen aus Alu-Rohr mit 8mm Durchmesser.

Abb. 6: QY21308XL4 Anordnung

Leistungsdaten & Vergleichstabelle & Abmessungen

Tabelle 4 & 5 – entsprechend wie Tabelle 2 & 3. **Tabelle 6:** Ele. aus 8mm Alu; ohne Boomkorrektur.

Abb. 7, 8, 9: Az/Elev.-Diagramme und SWR-Plot **Abb. 10, 11, 12:** QY21105x14 bei S52FO

Hybrid LFA Yagi Oblong Antennas

by Slobodan Bukvic, YU7XL

Is there anyone who did not hear IK3MAC via EME? Well, his 24x23 ele monster antenna system for 144MHz really does the job! There are a few photos of his antenna system available on the Internet. You can see that the antennas are a combination of yagi and loop elements, but no details are available. I started investigating this secret. My first step was to take EZNEC files of some good yagi designs, convert them into hybrid antennas and look at the simulations. I chose some of my 200Ω LFA yagis and also some YU7EF yagis converted to LFAs. First, with only the last director changed, a better gain and G/T was achieved. The feed impedance remained at 200Ω. Every subsequent director converted gave better and better results. Well, not quite! When this conversion process got to the elements close to the radiator, the antenna loss temperature started to increase, and despite the fact that the gain was a bit higher, the overall G/T ratio decreased. At the same time the impedance became unacceptable, so some changes in the element position had to be made, which made the G/T even worse. The best solution I found was to keep the reflector, radiator and first two to five directors as yagi elements and change the rest to loop elements. And... this is what the IK3MAC design approximately looks like! The antenna outperforms all existing ordinary yagi or loop antennas!

I do not know if Graziano will publish his design. I have found a loop rectangle with a width/height ratio = 1 gave the best results. This ratio gives an additional benefit, as you will see later. Probably the IK3MAC round loop is still better, but I find it is practically impossible to make a round loop model for EZNEC. I have to say that I have modelled several antennas for 50Ω feed impedance. The performance achieved was not as good as with 200Ω. Therefore all my models unless otherwise stated, are designed for 200Ω impedance and the feed point is at DE2. That means that a 1/2 coaxial balun is required for matching and balancing. Here are two of my most successful hybrid models:

Antenna QY21105x14

This antenna consists of 7 yagi and 4 oblong elements, made of Al rod 5mm diameter:

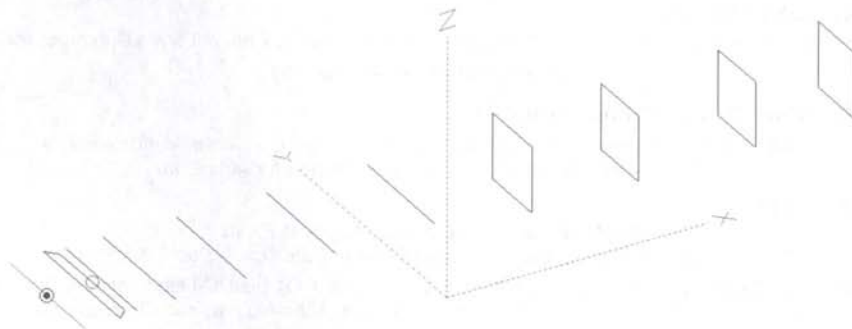


Fig. 1: QY21105x14 layout

Performance

Ant. type	No of ele	Length (m)	Gain (dBi)	F/B (dB)	F/Sh (dB)	F/Sv (dB)	H lobe (°)	V lobe (°)	TA (K)*	G/T (dB)*	Centre f (MHz)	Δf for SWR=1.5	
												Low-er	Upper
QY 2110 5XL4	11	6.0	15.4	28.3	19.0	18.3	33	36	220.4	-8.00	144.5	143.2	146.1
			15.3	28.2	18.9	18.2	33	36	222.5	-8.14			

Table 1 - The first row gives free space data with no losses, the lower row gives data with losses for aluminum included.

Comparison table

Type	Length (mm)	Gain (dBi)	TA	G/T	Freq (MHz) for SWR=1.5
QY21105XL4	6000	15.43	220.4	-8.00	143.200 – 146.100
		15.33	222.5	-8.14	
EF0211B-5	5980	15.13	229.1	-8.47	140.600 – 145.210
		15.03	230.8	-8.60	
G0KSC 12ELKKA	6048	15.22	228.7	-8.37	142.500 – 145.470
		15.14	229.8	-8.47	

Table 2 - The first row gives free space data with no losses
the lower row gives data with losses for aluminum included.

Dimensions

	RE	DE1	DE2	D1	D2	D3	D4	D5	DQ1	DQ2	DQ3	DQ4
Position	0	265	345	415	705	1250	1934	2695	3535	4355	5245	6000
Length	1018	1018	850	946	958	932	916	900	522	500	492	476

Table 3 - All elements made of Al rods Ø5 mm.

The height of oblong elements 500 mm, no boom correction included.

As you can see, the reflector and the first director are yagi elements. The directors DQ1 to DQ4 are oblong elements, 500 mm high, and their horizontal sizes are given in the table. Material for all elements is 5 mm diameter aluminium rod.

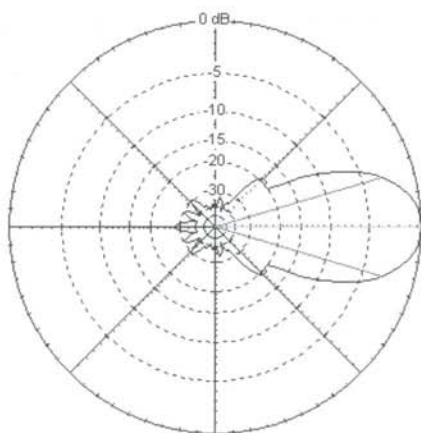
Simulation Results (which take account of aluminium losses)

Total Field

EZNEC+

Total Field

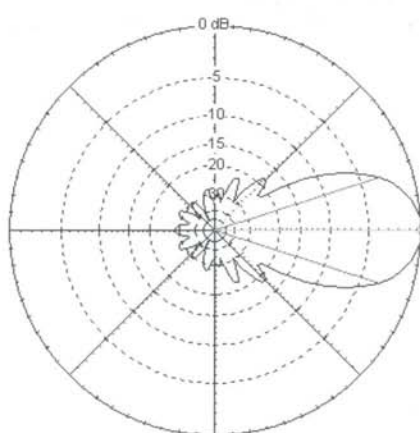
EZNEC+



144.5 MHz

Azimuth Plot
 Elevation Angle 0.0 deg.
 Outer Ring 15.33 dBi
 3D Max Gain 15.33 dBi
 Slice Max Gain 15.33 dBi @ Az Angle = 0.0 deg.
 Front/Back 28.22 dB
 Beamwidth 33.2 deg.; -3dB @ 343.4, 16.6 deg.
 Sidelobe Gain -3.57 dBi @ Az Angle = 45.0 deg.
 Front/Sidelobe 18.9 dB

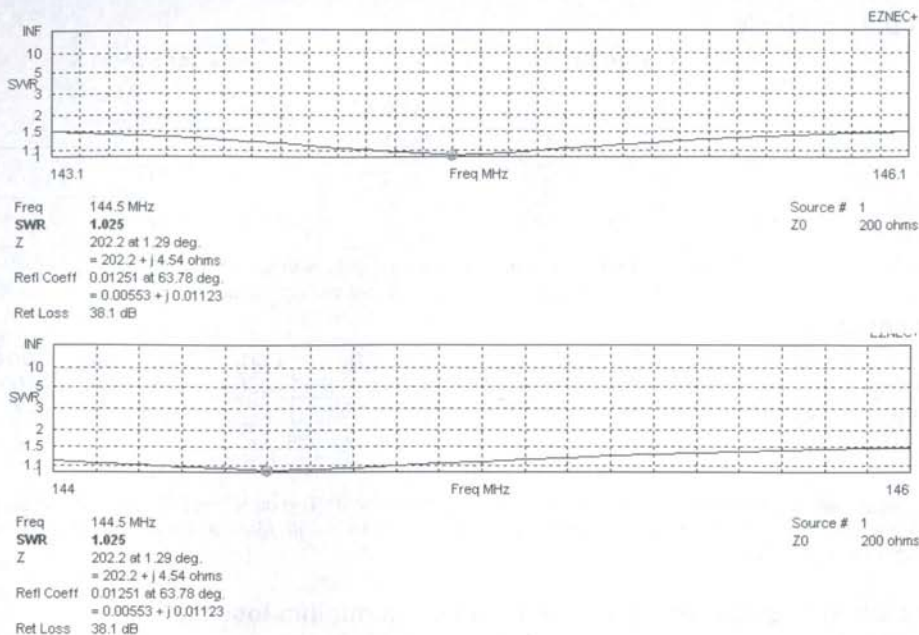
Fig. 3: Azimuth plot QY21105x14



144.5 MHz

Elevation Plot
 Azimuth Angle 0.0 deg.
 Outer Ring 15.33 dBi
 3D Max Gain 15.33 dBi
 Slice Max Gain 15.33 dBi @ Elev Angle = 0.0 deg.
 Front/Back 28.22 dB
 Beamwidth 35.6 deg.; -3dB @ 342.2, 17.8 deg.
 Sidelobe Gain -2.92 dBi @ Elev Angle = 45.0 deg.
 Front/Sidelobe 18.25 dB

Fig. 4: Elevation plot QY21105x14



Antenna QY21308XL4

This antenna consists of 7 yagi and 6 oblong elements, all made from 8 mm diameter Al tube:

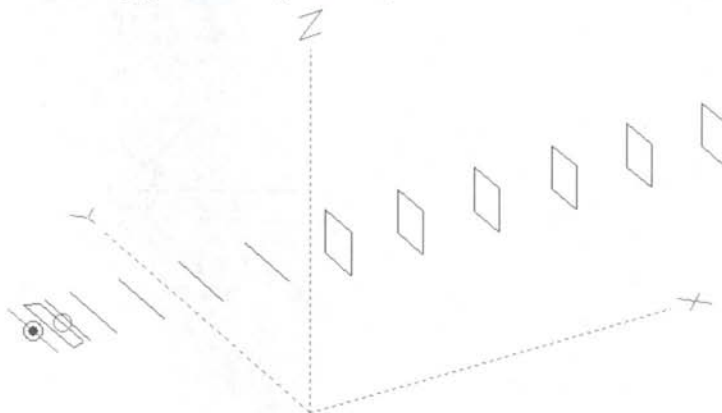


Fig. 6: QY21308XL4 layout

Performance

Ant. type	No of ele	Length (m)	Gain (dBi)	F/B (dB)	F/Sh (dB)	F/Sv (dB)	H lobe (°)	V lobe (°)	TA (K)*	G/T (dB)*	Centr. f (MHz)	Δf for SWR=1.5	
												Lower	Upper
QY21308XL4	13	7845	16.4	31.1	28.3	18.4	30.2	31.4	217.7	-7.01	144.5	143.2	146.2
			16.3	31.4	28.3	18.4	30.2	31.2	219.5	-7.14			

Table 4 - The first row gives free space data with no losses, the lower row gives data with aluminum losses included.

Comparison table

Type	Length (mm)	Gain (dBi)	TA	G/T	Freq (MHz) for SWR=1.5
QY21308XL4	7845	16.37	217.7	-7.01	143.200 – 146.200
		16.28	219.5	-7.14	
G0KSC 13ELKKB	7855	16.13	229.1	-7.28	142.650 – 145.370
		16.03	221.1	-7.42	
EF0213-5	8155	16.40	224.8	-7.12	143.660 – 144.620
		16.25	227.6	-7.32	

Table 5 - The first row gives free space data with no losses, the lower row gives data with aluminum losses included.

Dimensions

	RE	DE 1	DE 2	D1	D2	D3	D4	D5	DQ1	DQ2	DQ3	DQ4	DQ5	DQ6
Position	0	180	330	401	705	1251	1938	2705	3525	4355	5233	6128	6995	7845
Length	1012	990	868	942	951	925	907	887	526	504	496	486	476	456

Table 6 - All elements made of Ø8 mm Al rod. The height of oblong elements is 500 mm. No boom correction is included.

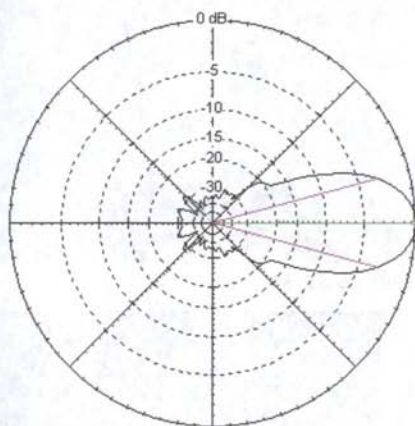
Simulation Results (which take account of aluminium losses)

Total Field

EZNEC+

Total Field

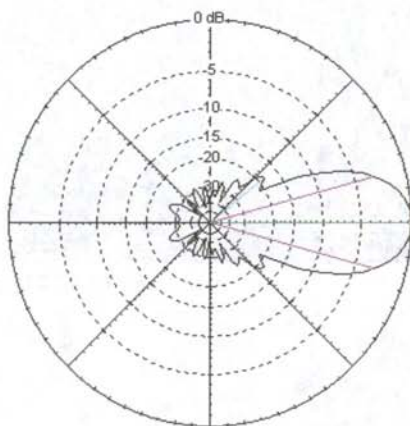
EZNEC+



144.5 MHz

Azimuth Plot
 Elevation Angle 0.0 deg.
 Outer Ring 16.28 dBi
 3D Max Gain 16.28 dBi
 Slice Max Gain 16.28 dBi @ Az Angle = 0.0 deg.
 Front/Back 31.43 dB
 Beamwidth 30.2 deg; -3dB @ 344.9, 15.1 deg.
 Sidelobe Gain -12.01 dBi @ Az Angle = 138.0 deg.
 Front/Sidelobe 28.29 dB

Fig. 7: Azimuth plot QY21308XL4



144.5 MHz

Elevation Plot
 Azimuth Angle 0.0 deg.
 Outer Ring 16.28 dBi
 3D Max Gain 16.28 dBi
 Slice Max Gain 16.28 dBi @ Elev Angle = 0.0 deg.
 Front/Back 31.43 dB
 Beamwidth 31.2 deg; -3dB @ 344.4, 15.6 deg.
 Sidelobe Gain -2.1 dBi @ Elev Angle = 39.0 deg.
 Front/Sidelobe 18.38 dB

Fig. 8: Elevation plot QY21308XL4

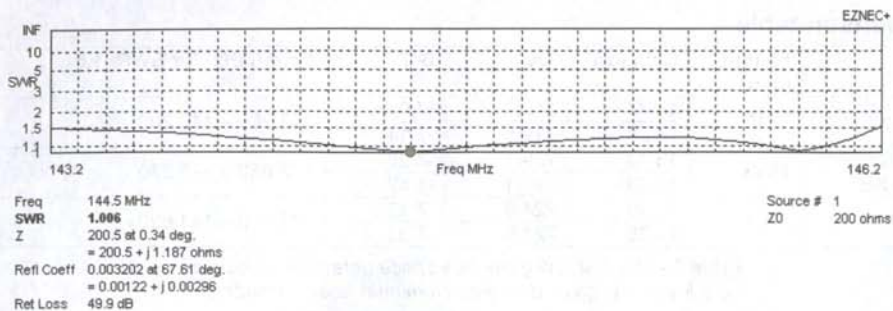


Fig. 9 SWR plot QY21308XL4 (simulation)



Fig. 10: QY21105x14 built by S52FO

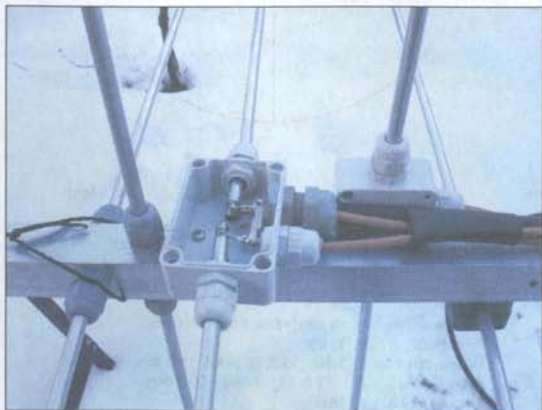


Fig. 11 & 12: Details of the driven elements

Making & mounting of WG flanges

by Philipp Prinz, DL2AM

Following my recent DUBUS articles about the construction of transverters with mechanical changeover [1, 2] there took place many discussions about the problem how to make the necessary waveguides and flanges. Professional flanges with rectangular holes are not available for most amateurs. In the meantime I have built five of these transverters with mechanical changeover (Figures 1 and 2) and have collected some experience that I can now share. I already gave a lecture about this topic at the last GHz Meeting in Tri Studne, Czech Republic.

Interesting details about dimensions and data of modules and parts which we need for our extraordinary hobby can be found on the website of Quinstar [3]. Also details about WG parts are there, and for example I found a transition from rectangular to circular WG that had a surprisingly low loss of only 0.7dB from 60 to 90 GHz.

For the DIY WG project described here, I used brass rod material of 20mm diameter from Conrad Elektronik [4] and of course a lathe. Flanges with a milled nose are very convenient for our purposes, see Fig. 2a. All of my flanges for the higher microwave bands are 20mm diameter and 4mm thickness, similar to the professional WG387 flanges which have 4 UNC 4-40 threads for WR19 to WR4 in the vertical and horizontal axis. These flanges have another two 1.5mm drillings at a 45° offset for the locating pins, which can be made from steel wire (also available from Conrad Elektronik) and the mating flanges have 1.6mm holes. If a professional WG387 flange is available and the locating pins are removed, the 8 drilling positions can be copied directly to the home made version. This technique helps to get more accurate measurement results on occasions when professional measurement equipment, WG transitions and WG387 to WG383 adapters are available. The professional UNC 4-40 connecting screws have a recessed head for use with a 5/64 inch or 3/32 inch hexagon wrench. Because UNC 4-40 is not easy to find in Europe, I used metric 3mm (M3) threads and hardware. The flanges that are used for mounting the dish to the transverter have four 3mm holes with 45° offset, like the well known Procom dishes. The flange of the transverter has four corresponding holes with M3 threads. This results in a system that allows all dishes to be exchanged. If the 1.5mm fixing pins are used on the flanges, the necessary drillings have to be perfectly exact.

These same 20mm flanges can also be used with the smaller WR19, WR12 and WR4 waveguides, if the following hints are considered. The diagonal length for WR19 is 7.75mm, for WR12 5.95mm, for WR8 4.15mm and for WR4 3.0mm. For example, if the centre of the flange is drilled 5.9mm for WR12 and the rear rim of the hole is slightly chamfered with a taper reamer, the WG can then be pressed into the flange until it is flush with the front side as shown in Figure 3. Instead of the taper reamer a universal countersink (or 'de-burring tool') could also be used (Figure 4). But first the WG itself has to be milled or filed square at the front side. This will be difficult without a lathe but an experienced engineer may do the job with a file, finishing by carefully cleaning up the WG opening with a scalpel.

When pressing the WG into the flange, the copper at the edges must not be pushed back, as otherwise the flange will not be inserted squarely into the WG and also will not fit tightly, both of which are very important. However, it is possible to file a slight chamfer on the four edges of the WG in order to fit the diameter of the hole in the flange, as shown in Figure 5. When the WG with the flange is put into a calliper, one can then see if the drillings are at the correct angles relative to the faces of the WG, and the flange can be rotated if necessary. Then the flange should be checked again for squareness (e.g. by rotating it in a lathe chuck as shown in Figure 6). Before soldering, the exposed end faces of the WG need to be protected with a heatproof paint to prevent solder from flowing inside the WG (Figures 7, 8 and 9). For soldering, the WG is set with the flange to a plane surface, preferably a ceramic tile (Figure 10) to prevent heat loss and so that the solder will flow flush with the end of the WG. Then the flange and WG are heated with a gas burner until the solder melts and fills the entire drilling hole. Now one end is complete. If a second flange has to be soldered on to the other end of the WG, the same procedure has to be repeated there. But before the second flange is heated, a copper plate or similar should be placed on the upper (completed) flange in order to prevent the solder melting again (Figure 10).

So now we have a WR 12 waveguide with similar flanges like the WG387 (Figure 11). If a transition from rectangular WG to circular WG is needed, I use a conical milling cutter (Figure 12) and then cut back the

rectangular WG until the opening is circular, flush and square to the axis. In this way flanges can be mounted to waveguides including WR19 and smaller. Larger WG than WR19 with flanges similar to WG387 can not be made this way.

Addenda

Cleaning waveguides should not be done with cotton, but with a piece of linen that is fixed to a wire loop (Figure 13). The linen should be soaked with acetone solvent and pulled through the WG. Use only linen; cotton always leaves fluff which causes losses.

When measuring power with a thermal head it will be of advantage to use an isolator with known loss, as it has bandpass properties which reduce the effects of harmonics upon the measurement. I could clearly notice this bandpass characteristic in my measurements; see Figure 14.

I have also built three harmonic mixers and one subharmonic mixer for 122GHz like I had described in DUBUS 4/2006 [6]. These use the zero bias diode(s) HSCH 9161. I have made signal to noise measurements with my elaborate measurement equipment, using two 50dB WR10 attenuators with 0.1 dB reading (Figure 15). For comparison a homemade harmonic mixer with the single diode MA4E1317 was used. Using my measurement method, I regretfully can not confirm the good measurement results which were published by Jürgen DC0DA in DUBUS 2/2010 [7].

Final

I know that mechanical work is not everybody's passion, but in this case precision is really necessary. Greetings from Philipp DL2AM, who really *does* enjoy mechanical work!

References & Sources

- [1] Philipp Prinz, DL2AM: A 76GHz Transverter – with mechanical changeover, DUBUS 3/2009 p.9
- [2] Philipp Prinz, DL2AM: A 47GHz Transverter – with mechanical changeover, DUBUS 1/2010 p.32
- [3] www.quinstar.com
- [4] www.conrad.de
- [5] www.selva.de
- [6] Philipp Prinz, DL2AM: 122 GHz Transverter with new multiplier, DUBUS 2/2006, pp.8
- [7] Jürgen Dahms, DC0DA: Compact 122 GHz / 134 GHz mixer head for direct dish mounting, DUBUS 3/2010, p.76



Fig. 1: Parts of two 76 GHz transverters with mechanical changeover



Fig. 2: Completed transverters for 47 GHz and 76 GHz with mechanical changeover



Fig. 2a: Flange with milled nose



Fig. 3: DIY drilled flanges for WR 19, WR 12 and WR 4, with a professional WG387 flange



Fig. 4: Tapered reamers, universal countersink and UNC 4/40 taps



Fig. 5: Checking angular positioning with a calliper



Fig. 6: Checking a WG & flange for squareness



Fig. 7: WR 19 and WR 12 waveguides with flange. The exposed ends of the WG have been painted with a black solder resist before soldering.



Fig. 8: like fig. 7 plus professional flange WG387



Fig. 9: WR 19 and WR 12 waveguides pressed into flanges ready for soldering



Fig. 10: Upper flange is already soldered, and now protected from melting by a copper plate



Fig. 12: Universal countersinks and conical milling cutters for circular WG adapters



Fig. 11: Completed WGs with flanges, similar to WG387 and WG383, with flange connecting screws and screwdrivers for inner hexagon



Fig. 13: Waveguide cleaning set

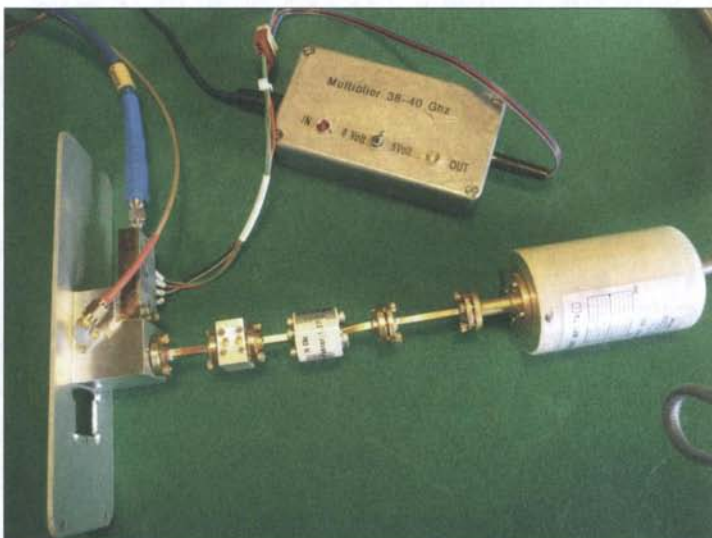


Fig. 14: Measurement setup with 76GHz filter (loss <3db) and isolator



Fig. 15: Measurement setup for 122 GHz with two 50 dB attenuators reading to 1/10 dB

Test beacon with passive BAT15-099 Diode Multiplier for 47GHz - also for 24, 76, 122 and 241GHz

by Wolfgang Demmer, DD8BD

My intention was to make a homebrew multiplier, with a WG flange for a horn antenna, that I could use as a 47GHz test beacon. However, my workshop and skills were not enough to build the design I had developed. So I asked Bert, PE1RKI [1], if he could build this multiplier for me and he agreed to attempt this difficult task.

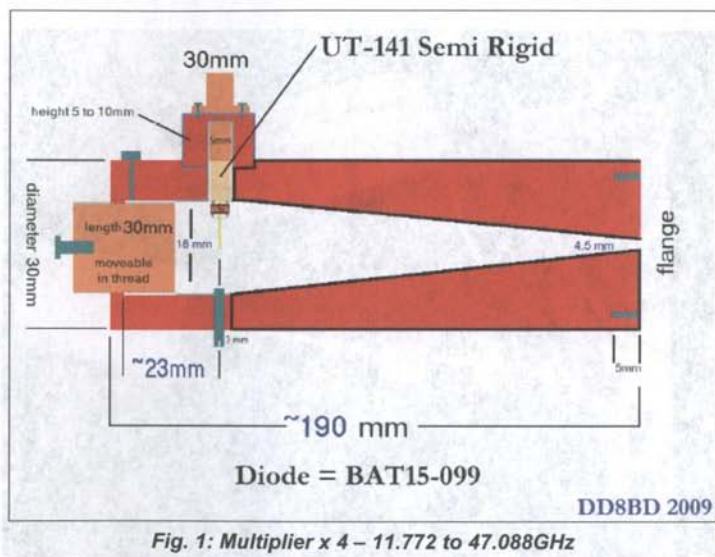


Fig. 1: Multiplier x 4 - 11.772 to 47.088GHz

As you can see from the sketch in fig. 1, this multiplier consists of a conical WG, which continuously gets smaller in diameter towards the output. My intention was to make a gentle transition from the input frequency up to 47GHz with the help of this conical form, which would have only small mismatches compared to stepped circular WG. It would also have better high pass filter properties. Another piece of straight circular WG, attached by a flange to the output of the conical WG, could make an additional high pass filter if needed. For optimum matching of the input signal to the WG there is a tunable short circuit. The diameter at the output end of the conical WG put is 4.5mm and suits the circular WGs which I use for 47GHz.

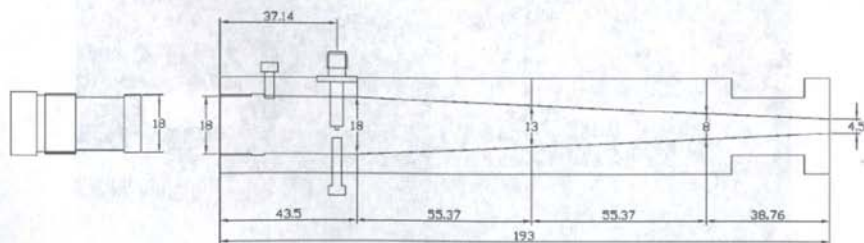


Fig. 2: Technical drawing of the multiplier from PE1RKI.

After a while I got a superb technical drawing from Bert, that showed how he would build my idea. See fig. 2. He suggested milling four different separate parts which would then be screwed together to form the complete assembly. Finally Bert built a beautiful mechanical masterpiece.



Fig. 3: Completed milled multiplier

Figure 3 shows the completed multiplier with a knurled fine thread screw for the adjustable short circuit. This has a 0.5mm pitch thread that allows easy adjustment. In this configuration I was able to get good beacon signals even on 241GHz, when driven with 100-120mW (as much as the diode will stand).

When the final circular WG section of the multiplier is removed, the output diameter of the previous section is 8mm. From here one could get a harmonic signal on 24GHz using a suitable LO signal at the input. For such "low" frequencies one needs a significantly lower LO level.

The conclusion is that, depending on the input frequency and level, which could be increased by an additional amplifier, this multiplier is usable as a test beacon source on 24, 47, 76, 122 and 241GHz. I recommended using conical horn antennas at the output.

If a stable signal generator or synthesizer like the HP8671B (2-18GHz) is available, one needs no special frequency generating chain for each mm band, but can feed the synthesizer signal to the input of the multiplier. For the protection of the synthesizer I recommended using an isolator. Figure 4 shows how to connect the BAT diode to a piece of semi rigid UT141 coaxial cable that had been soldered to a SMA flange jack.

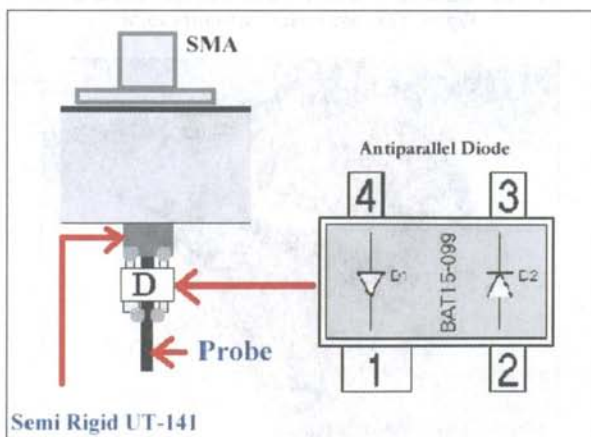


Fig. 4: Installation of the BAT15-099 diode for the 47/76/122/241GHz multiplier

Pins 3 and 4 of the diode have to be soldered to the outer shield of the UT141 cable and pins 1 and 2 to the inner conductor. As it is an antiparallel diode, connecting the other way around is also possible, of course. During my tests it was clearly noticeable that the multiplier delivered the strongest beacon signal if both diodes of the BAT15-099 were connected. When just one of the diodes was connected significantly weaker signal levels were observed. Now follow a few more photos of the multiplier.



Fig. 5: Open multiplier



Fig. 6: Test assembly without diode



Fig. 7: BAT15-099 – just one half is soldered here

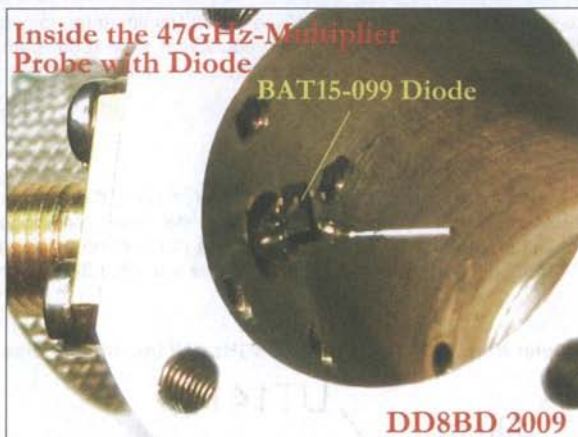


Fig. 8: BAT15-099 – both diodes soldered and SMA jack inserted



Fig. 9: Multiplier with conical horns for 47GHz and 76GHz

There is an additional straight piece of circular WG inserted as adapter between the multiplier and the horn which acts as a high pass filter. Depending on the inner diameter this can be used for different mm bands.

Results

When tuning the threaded short circuit screw, one finds both minima and maxima. When the cable to the multiplier or the input power is changed, a re-adjustment of the screw will improve the match and the output power will increase.

Tests on 47GHz with a DB6NT LO (11,772225GHz) showed that the highest output was obtained from the multiplier when the tuning screw was inserted almost completely into the cavity. For additional matching the 3mm tuning screw on the opposite side of the input coupling pin can be used, if it is not already used to hold something else like a varactor diode. A conical horn antenna like the one I described in DUBUS 3/2008 [2] can be attached to the multiplier with a flange. Of course, other antennas with a suitable flange or with an adapter can be used. I also have a conical 76GHz horn now with a suitable adapter, which can be attached to the multiplier. In order to estimate the performance of this multiplier I made some measurements on 47GHz over a distance of 1m, using a DB6NT LO on 11.772225GHz to drive the multiplier.

- I. Multiplier with conical horn without any diode: Signal level up to 33dB.
- II. Multiplier with conical horn with $\frac{1}{2}$ BAT15-099 diode: Signal level up to 47dB.
- III. Multiplier with conical horn with 2 connected halves of the BAT15-099: Signal level up to 62dB.
- IV. DB6NT-LO (without multiplier) with homemade broadband Vivaldi antenna: 40dB.

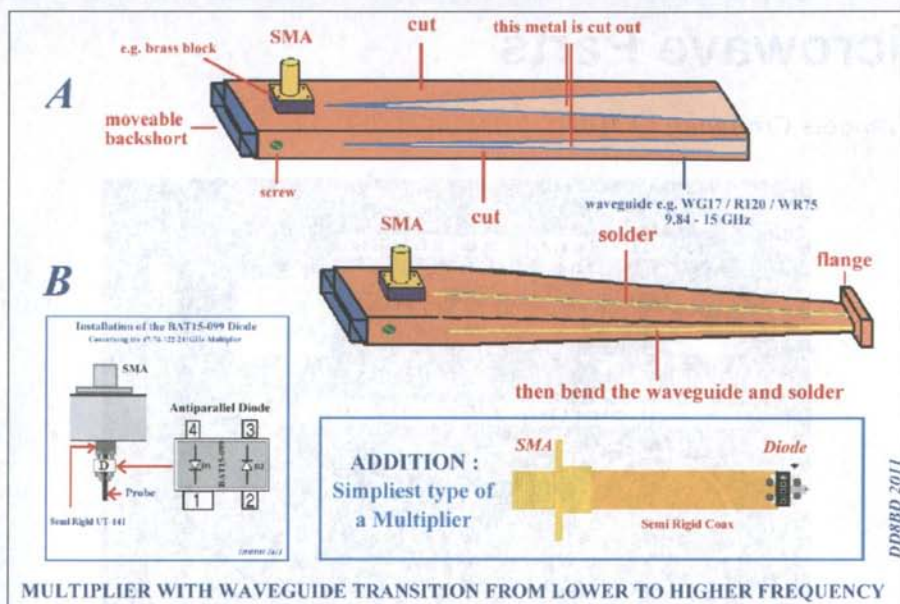


Fig. 11

The simplest solution to generate a signal on a mm band is shown as addition in figure 11. It consists of a piece of semi rigid coaxial cable, an SMA jack and a BAT15-099 diode. Such an assembly could be placed as a feed in a small dish, a parabolic shaped metal fruit bowl, a shaving mirror or even a soup ladle in order to get some antenna gain.

I hope that I have given you some inspiration for your own experiments and wish you a lot of success and fun! Vy 73 Wolfgang Demmer, DD8BD - E-Mail: dd8bd@gmx.net

Literature and Sources

[1] <http://members.chello.nl/b.modderman/custom.html> Email: PE1RKI@chello.nl

[2] W. Demmer, DD8BD: Conical Horn Antenna for 47GHz, DUBUS 3/2008 and DUBUS TECHNIK VIII



Fig. 12

Microwave Parts

by Francois Cronauer, LX1DU



Fig. 1: Top, from left: 1. Variable attenuator 60-90GHz. - 2. Noise generator 47-120GHz, 22dB ENR. 3. Detector 47-76GHz, 0dBm = 300mV. Below: 4. WG feed 47-76GHz 5. WG feed 76-122GHz. Right: Waveguides WR12 and WR15

Abb. 1: oben, von links nach rechts: 1. Abschwächer variabel 60-90 GHz. - 2. Noise-Generator 47-120 GHz, 22 dB ENR. 3. Detektor 47-76 GHz, 0 dBm=300 mV, darunter: 4. Krückstock 47-76 GHz. 5. Krückstock 76-122 GHz, rechts aussen: 6. Hohlleiter WR12 und WR15

LX1DU fabricates special and difficult to find parts for Microwave amateurs. Some of these parts are shown in fig. 1 and fig. 2 and briefly described in the following. If you need more information or if you need such parts, please contact LX1DU.

1. A variable attenuator for 60-90GHz is made from WG WR12 with UG387U flanges. As attenuating material 0,002 inch Mylar foil (NORSAL Industries Inc., pro nova Messtechnik GmbH) is used, which is inserted through a 0.1mm slot into the WG and enables a variable attenuation from 2 to 25dB.

2. A noise generator is made from WG WR12 or WR15 (length about 70 mm) with suiting flanges and a noise diode NC406 (Wireless Telecom Group), which works from 60-120 GHz with DC=25mA, ENR 20 dB (E.T.O.N measuring station, Noisemonitor 7310). Fig. 2 shows a noise generator for 76 GHz.

3. A detector for 47 or 76GHz is made from a milled casing with WG input for 47-80 GHz. Output is SMA. This detector is especially made for the PROCOM Wattmeter (MCW 3000, source Eisch Electronic, see DUBUS 4/2010, p. 3) but can be calibrated also for any other power meter (-30 to -20 dbm). The detector is a point contact diode with coaxial casing from Sylvania for 20-100 GHz. Sensitivity of the detector is about 300mV/mW (300 mV at 0 dbm).

4. WG feed for 47 and 76GHz, and 5. WG feed for 76 and 122GHz.

These WG feeds in 'walking stick' shape are made for PROCOM dishes with 112mm focus (source Eisch Electronic) but can be used also for other dishes depending on their focal length. These feeds are made from very thin WG material (0.5mm), which can be silver plated on request. Parabolic feeds from thin WG have significantly less sidelobes. LX1DU has constructed a special machine to build these waveguides for 47 to 500GHz with a length of up to 1000mm.

6. Waveguide material WR12 and WR15.

LX1DU builds also other microwave parts like transverters, beacons and filters.

Contact: LX1DU, Francois Cronauer, 140 rue Aessen, L-4411 Soleuvre, Luxembourg



Fig. 2 / Abb. 2: Noise generator for 76GHz / Rauschgenerator für 76 GHz

Deutsch: LX1DU baut spezielle schwierig zu beschaffende Mikrowellenteile, die in Abb. 1 und 2 gezeigt sind und im folgenden kurz beschrieben werden. Für mehr Informationen oder den Bezug bitte den Autor kontaktieren.

1. Der Abschwächer besteht aus einem Hohlleiter WR12 mit den Flanschen UG-387U. Als Dämpfungsmaterial wird 0,002 inch Mylarfolie (NORSAL Industries Inc., pro nova Messtechnik GmbH) verwendet. Diese wird über einen Schlitz von 0,1 mm im Hohlleiter eingeschoben und ermöglicht die variable Einstellung der Dämpfung (2-25 dB).

2. Der Noise-Generator besteht aus Hohlleiter WR12 bzw. WR15 (Länge ca. 70 mm) mit den passenden Anschlussflanschen. Als Rauschdiode findet eine NC406 (Wireless Telecom Group) Verwendung, 60-120 GHz, DC=25mA, ENR 20 dB (Messplatz E.T.O.N, Noisemonitor 7310). Abb. 2 zeigt einen Rauschgenerator für 76 GHz.

3. Messdetektor 47 oder 76 GHz: Gefrästes Gehäuse, Eingang Hohlleiter 47-80 GHz. Ausgang über SMA-Buchse. Der Detektor ist speziell für das PROCOM Wattmeter (MCW 3000, Bezug Eisch Electronic, siehe DUBUS 4/2010, Seite 3) konzipiert, kann aber für jedes andere Wattmeter kalibriert werden (-30 bis -20 dBm). Als Detektordiode dient eine Whiskerdiode im Koaxgehäuse (Sylvania), 20-100 GHz. Empfindlichkeit des Detektors: ca. 300mV/mW (300 mV/0 dbm)

4. Krückstock 47 und 76 GHz und 5. Krückstock 76 und 122 GHz

Die abgebildeten Krückstöcke (Erreger) sind für PROCOM-Spiegel, Focus 112 mm (Vertrieb Eisch Electronic) gedacht, können jedoch je nach Länge für jeden anderen Spiegel verwendet werden. Der Krückstock besteht aus einem sehr dünnen Hohlleiter (0,5 mm), welcher auf Wunsch versilbert wird.

Parabolerrger aus dünnen Hohlleitern haben viel weniger Nebenzipfel. Der Verfasser baute speziell für diese Hohlleiter ein Gerät, mit dem Hohlleiter für 47-500 GHz bis zu einer Länge von maximal 1000 mm gefertigt werden können.

6. Hohlleiter WR12 und WR15.

Der Verfasser baut außer den oben beschriebenen Teilen auch Transverter-Baken und Filter im GHz-Bereich.

A Low Loss x4 Subharmonic Mixer for 24GHz

by Paul Drexler, W2PED

Introduction

At low microwave frequencies, LO generation typically requires using a low frequency crystal oscillator or synthesizer, and then multiplying up to the final LO frequency. Low cost multipliers work well to about 10GHz, but at 24GHz and above, many of the low cost MMIC components simply don't work well; generating a fundamental mode LO for the 24GHz band is somewhat difficult.

Another approach (as described here) is to use a subharmonic mixer. This type of mixer uses an LO input lower than the normal LO input frequency and the mixer's transfer function effectively multiplies the LO input frequency by n within the mixer to get to the desired high frequency. A subharmonic mixer makes for simpler LO hardware; this is especially helpful in getting on a higher frequency band like 24GHz. A higher order ($n=4$) subharmonic mixer is presented here for the 24GHz amateur band (see figure 1 below). The conversion loss is only slightly higher than that of a fundamental mode mixer, and only a ~6GHz LO is necessary for operation at 24GHz. Figure 2 presents several frequency schemes for subharmonic mixers for the 24048MHz and the 24192MHz frequency bands.

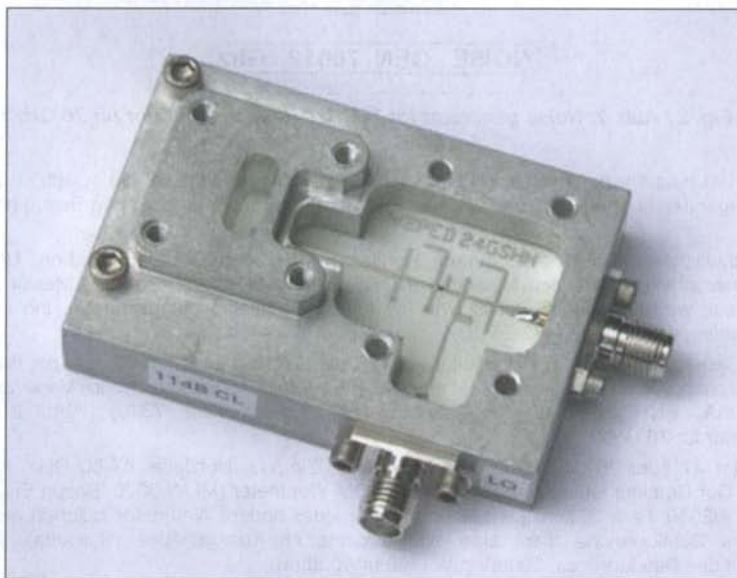


Figure 1: W2PED 24GHz subharmonic mixer

RF Freq	IF Freq	LO Input
24,048	144	5976
24,048	432	5904
24,192	144	6012
24,192	432	5940

Figure 2: Several frequency mixing schemes for 24GHz

Characteristics of Subharmonic Mixers

A subharmonic type mixer allows us to use a simpler LO, but requires a more difficult mixer design. Also, the efficiency of subharmonic mixers can sometimes be very low; for example, the Hewlett Packard HP 11970A subharmonic mixer has a conversion loss of nearly 20 dB! Some of this is because the mentioned mixer is very broadband. Fortunately, most amateur work uses narrow band operation, and over narrow operating bands the circuit efficiency can be much improved. Published narrowband x2 ($n=2$) type subharmonic mixers for 24GHz have a typical conversion loss of 8-10 dB, so we would expect more loss for a x4 ($n=4$) subharmonic mixer. In addition to higher conversion loss, a subharmonic mixer may have more spurious signals due to the internal multiplication process. For some broadband commercial applications this may be an issue, but for amateur applications this is not a problem.

Subharmonic mixer work by others

It's important to note that a number of amateurs have published work on subharmonic mixers. Jim Dietrich WA0RDX published a novel x2 subharmonic mixer design for 1296MHz in the October 1978 issue of Ham Radio magazine. [1] His mixer uses two back-to-back diodes in series with a microstrip transmission line resonant circuit. The DB6NT subharmonic mixer for 24GHz (and other bands) is another example of a x2 subharmonic mixer. [2] Brian Justin WA1ZMS has made good use of a waveguide type subharmonic mixer for his 145 and 241GHz work [3] and in fact has set several amateur microwave DX records with this hardware. Brian modified a x3 multiplier for use as a subharmonic mixer by adding an IF lowpass filter and biasing the diodes. Brian also published a nice discussion of various diode multiplier and mixer types in the 2005 Microwave Update Proceedings. [4] This work focused on a 24GHz subharmonic mixer design using microstrip. Other waveguide-based subharmonic mixer designs exist and numerous articles have been published in DUBUS and other publications. I have not done an exhaustive search on this subject but one such article uses a WG based subharmonic mixer for 47GHz by Peter Riml OE9PMJ. [5]

A higher order mixer

I thought it might be useful to design a microstrip based subharmonic mixer for 24GHz that would use a x4 LO scheme ($n=4$ in the above discussion). My thought was that, if successful, this would make a 24GHz transverter somewhat easier to build by requiring only a ~6GHz LO for the typical 144MHz IF scheme. I feel it's easier to generate a 6GHz LO signal than it is to generate a 12 or 24GHz LO. So this scheme has the advantage of using a lower frequency LO, but its disadvantages are that it's a more complex design, and will have more conversion loss than a fundamental mode mixer. The most promising mixer topology that I found for a x4 subharmonic mixer design (Figure 3) is based on work described in the IEEE literature by A. Madjar [6].

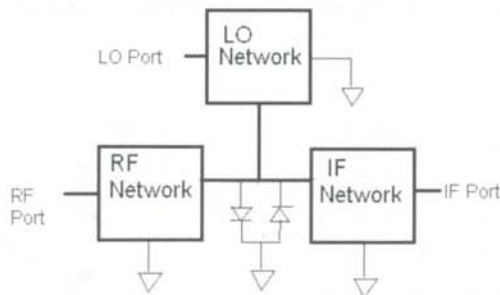


Figure 3: Subharmonic mixer block diagram

There are several key things that allow the mixer circuit work as an efficient subharmonic mixer. The first is that two diodes are used, placed back-to-back as shown above. This so-called "anti-parallel" arrangement suppresses circulating currents at the LO input frequency and its even-numbered harmonics ($n=2, 4$, etc) while at the same time promoting mixing with those same even-numbered harmonics. This allows a properly designed mixer to operate with low conversion loss, while at the same time providing substantial suppression of some major LO harmonics. Both of these features are quite contrary to normal amateur expectations! Secondly, note that there are filters used at the RF, IF, and LO ports of the mixer. Each filter is designed to pass the desired signal (RF, LO or IF) while being reflective to other signals that are internally generated within the mixer. The filters are discussed in more detail below.

Like most high frequency circuits, best results are obtained by keeping all parasitic inductances and capacitances as low as possible. For 24GHz this means using *very* small components. For a mixer, it's also important to use diodes that are well matched in their DC characteristics. Fortunately, some semiconductor manufacturers have made matched pairs of high frequency schottky diodes available at reasonably low cost. One such anti-parallel diode pair is the M/A Com MA4E1318, shown in Figure 4 below. Note that the parasitic capacitance is less than 0.06pF!

The good news here is that diodes with good performance to high frequencies are offered in conventional surface-mount type packages, so that no wire-bonding is required. A conductive epoxy can be used to attach the diodes to the mixer printed circuit board. The bad news is in figure 4-B: these devices are only 0.66 mm (0.026 inches) in length.... really small!! A good microscope and a very steady hand are required to work with these components.

MA4E1317, MA4E1318, MA4E1319-1, MA4E1319-2, MA4E2160



GaAs Flip Chip Schottky Barrier Diodes

M/A-COM Products
Rev. V3

Electrical Specifications @ + 25 °C

Parameters and Test Conditions	Symbol	Units	MA4E1317			MA4E1318		
			Min.	Typ.	Max.	Min.	Typ.	Max.
Junction Capacitance at 0V at 1 MHz	C _j	pF		.020			.020*	
Total Capacitance at 0V at 1 MHz*	C _T	pF	.030	.045	.060	.030*	.045*	.060*
Junction Capacitance Difference	ΔC _j	pF					.005	.010
Series Resistance at +10mA*	R _s	Ohms		4	7		4	7
Forward Voltage at +1mA	V _{F1}	Volts	.60	.70	.80	.60	.70	.80
Forward Voltage Difference at 1mA	ΔV _{F1}	Volts					.005	.010
Reverse Breakdown voltage at -10uA	V _{br}	Volts	4.5	7				
SSB Noise Figure	NF	dB		6.5*			6.5*	

Figure 4-A: MA/COM single and anti-parallel mixer diode specifications

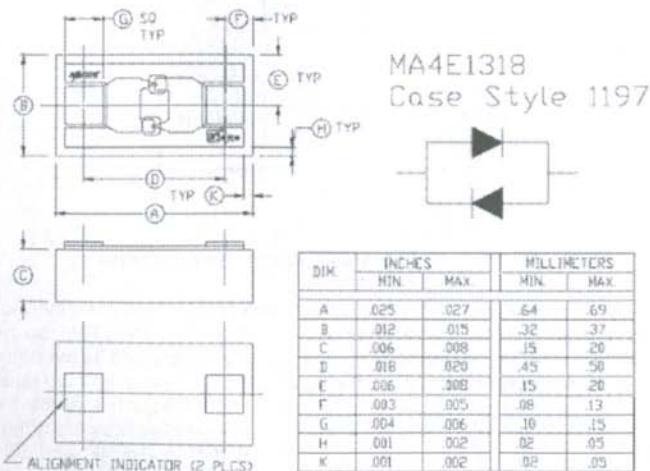


Figure 4-B: MA/COM anti-parallel mixer diode package

Mixer theory of operation

The mixer consists of the anti-parallel diodes, the RF filter, LO filter, and IF filter network (see block diagram in figure 3). The mixer operation can be described as either a receiver mixer and a transmit mixer; we'll refer to it as a receiver mixer to keep things straight. It's also less confusing to refer to the various frequencies as approximate numbers. Starting with the LO input, an external 6GHz local oscillator signal (~ 10 dBm) is fed to the LO input connector; a narrowband edge-coupled band pass filter is centered at 6GHz. It filters the LO energy, but more importantly, it is reflective to all other frequencies. When the 6GHz LO energy hits the diode pair, LO harmonics are internally generated; the 6GHz $\times 4$ energy becomes our desired high frequency LO within the mixer. Additional LO harmonics are generated at the $\times 2$, $\times 3$, $\times 5$ and higher multiples, but all of these signals are essentially reflected back towards the mixer. With optimization, the reflected energy improves the mixer efficiency, which is essential achieving a low conversion loss.

The RF network has two purposes. First, it works as a low loss, wideband bandpass filter centered at 24GHz. The BPF provides some degree of out-of-band rejection, but more importantly it keeps the various LO signals from radiating out the RF input of the mixer. It also looks reflective to the LO energy which is reflected back towards the diodes. Lastly, the IF section of the mixer passes the mixer's IF output signals, and just as in the other portions of the mixer, it also reflects LO energy back toward the mixer diodes. This is done using high-Q *idler circuits* (open circuited microstrip stubs). Each idler circuit is resonant at its own required frequency, so when you look at the final artwork for the mixer, you may agree that the IF section is not simple. Each of the main sections of the mixer is now discussed in greater detail as follows.

RF filter section

The RF bandpass filter started as a simple arrangement of two short-circuited $\lambda/4$ lines forming resonators at 24GHz. The resonators are direct-coupled through another transmission line, which again is approximately $\lambda/4$ in length. All lines are $\sim 50\Omega$. The beauty of this filter topology is that no tightly coupled lines are necessary, as they would be in an edge-coupled filter, so the typical physical layout is quite tolerant. The resulting filter response is a broad bandwidth bandpass at 24GHz.

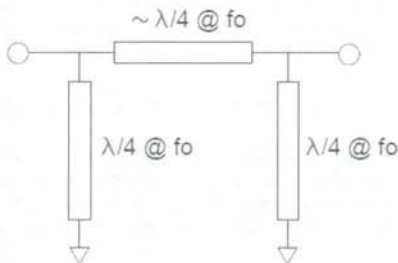


Figure 5: Basic RF filter

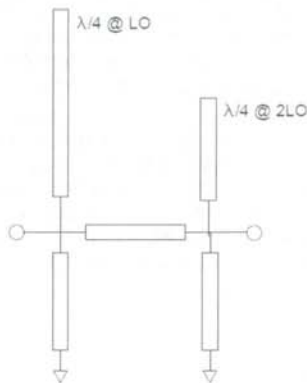


Figure 6: RF input filter with LO harmonic stubs added

Next, resonant stubs were added to the filter at the input and output to provide a short circuit response at the LO and $2\times LO$ frequencies. Each of the open circuited microstrip stubs are $\sim \lambda/4$ in length at it's designed frequency. These look like trap filters at the LO and $2\times LO$ frequencies.

The RF filter was first designed using rough hand calculations and then optimized using the student version of the Ansoft Designer circuit design software [7]. Line lengths were adjusted for best match and insertion loss at 24GHz, and best resonance of each trap. The predicted RF filter response is shown in figure 7. The Y1 (S21) trace shows that a low insertion loss (0.25 dB) is predicted at 24GHz, with the LO and $2\times LO$ frequencies being highly attenuated (reflected). The Y21 (S11) trace without markers shows the return loss at each frequency. The RF filter is designed to have a very good return loss at 24GHz, and a reflective impedance at the LO and $2\times LO$ frequencies. The $3\lambda/4$ resonance of the LO stub provides an additional null at the $3\times LO$ frequency (roughly 18GHz), so for two added stubs we get three nulls for the 6GHz LO and it's major low-order harmonics.

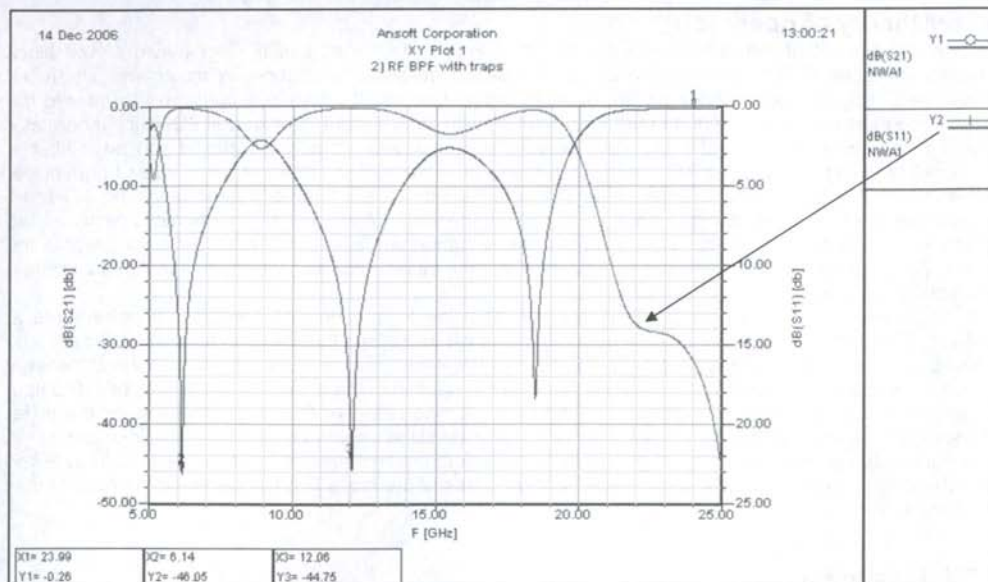


Figure 7: Predicted response of RF filter with stubs

LO filter section

The LO bandpass filter is a simple edge-coupled design made from two $\lambda/4$ lines at the 6GHz LO frequency. A more complex filter is unnecessary as the LO filter just has to pass the LO while being reflective at all other frequencies. The centre frequency was designed for 6GHz and the bandwidth was made wide enough to cover other desirable LO frequencies in this region. The wider bandwidth also makes sure that variations in the board etch process don't spoil the performance. The coupled line portion of the filter is approximately 0.4 inches (10mm) in length.

In order to present a 50Ω impedance match at the filter input/output, the coupled transmission lines need to be a higher impedance and are thus very narrow.

Using 10 mil (0.25mm) Teflon type board material, the lines are ideally 4 mils (0.1mm) wide with 4 mil spacing, but these small dimensions are difficult to etch reliably, even in commercial manufacturing processes. To make the filter more reproducible, I chose to compromise the design a bit and made the coupled section lines wider and slightly farther apart. This degrades the impedance match, but stubs at the input and output correct the problem.

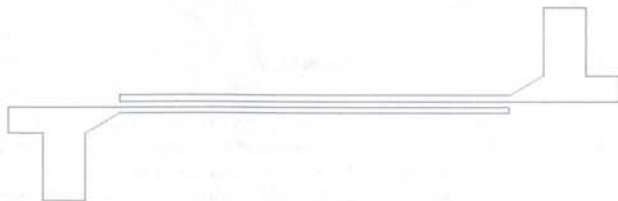


Figure 8: Layout of edge coupled LO filter

The design of edge coupled filters can be especially difficult unless the design includes all of the coupling effects. The best way to do this is to use an electromagnetic (EM) simulator. An EM simulation showed that the first design was shifted in frequency and would have given about a 3dB insertion loss. The desired centre frequency was obtained by adjusting the filter length and then re-analyzing using the EM simulator. The predicted insertion loss of the LO filter is less than 0.5dB at 6GHz.

IF network

The IF section of the mixer might be thought of as the "low frequency" part of the design, but in fact it needs to do considerably more than that. The IF section injects the LO energy into the mixer diodes, passes the IF signals to/from the diodes, and reflects the RF, LO, and 3LO energy back toward the diodes. The RF energy and all LO products are reflected back toward the diodes for two reasons:

- To block undesired signals from the IF output (i.e. to attenuate the RF and LO signals)
- Reflecting all signals back toward the diodes (except the IF) gives low conversion loss.

Each of the open circuited microstrip stubs are $\sim \lambda/4$ at its respective design frequency, and these stubs can also be thought of as high-Q *idler circuits*. They are very much like the resonant stubs used in the 432 to 1296 varactor tripler circuits of the past. Best mixer performance is very sensitive to the stub lengths. If a stub is off in length by more than a small fraction of a millimeter, the mixer performance will quickly degrade. The highest-frequency stubs are closest to the diodes in order to keep the reflected transmission paths of the RF and 3LO energy short. All of the interconnecting line lengths were optimized for best performance.

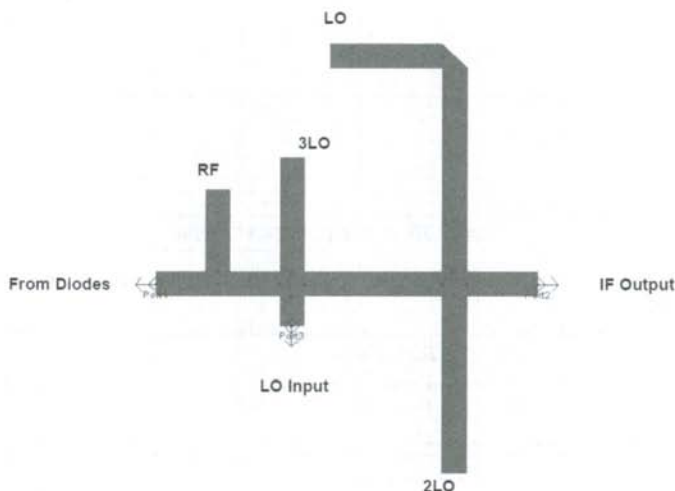


Figure 9: IF / LO injection network

Connecting it all together

The mixer was first designed for 10mil (0.25mm) thick Rogers 5880 high frequency circuit board material. The material was chosen for very good high frequency circuit properties including low dielectric losses (loss tangent ~ 0.0009 at 10GHz). A 10 mil board thickness was chosen to keep the transmission line widths a reasonable width for the frequency of operation. If 5 mil (0.127mm) material had been chosen instead, the LO bandpass filter would have required *extremely* narrow lines. Once the basic design of each section was completed, the next step was to connect it all together and analyse the circuit with the aid of Ansoft Designer, a nonlinear circuit simulator. An LO level of +10dBm was used. The first results were not very good... the predicted insertion loss was over 20dB! This was because of: 1) interactions between the sections; 2) each of the filter networks having been designed for an idealized 50 Ω load, when the diodes actually present a rather different impedance; and 3) since each of the stubs are high-Q, it was inevitable that some further optimization would be required. The next step was to optimize each stub and transmission line length to obtain the best performance. The goal was to make the mixer conversion loss as low as possible, while keeping the undesired signals as low as possible. Fortunately, as each of the undesired outputs is reduced, the mixer efficiency generally improves. Once satisfied with the initial circuit model, a more accurate electromagnetic (EM) analysis of the physical structure could then be made. As mentioned earlier, the use of EM analysis gives much more accurate results as it takes electromagnetic effects into account including mutual coupling between transmission lines. The downside is that this is a 'computationally demanding' exercise for the PC and it tends to be very time consuming. Another issue is that it's not easy to do EM analysis and optimization at the same time, so this makes it a multi step process. I decided that it would be worth the effort to go the extra step and to 'fine tune' the

design using the more accurate EM analysis; this would give the best chance of a first-time success when the prototype was constructed. Working at this greater level of detail, two slightly different mixers were designed – one for a 144 IF, and another for a 432 IF. Figure 10 illustrates the artwork layout of the mixer. The RF input is on the left, LO input at the bottom trace, and the IF is on the right. The mixer length is just over 0.75 inches. The LO filter appears to be one solid line unless the image is magnified.

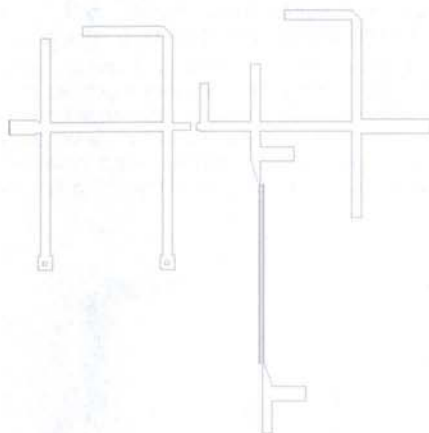


Figure 10: Prototype mixer layout

The plot in figure 11 shows the predicted spectral response of the mixer (432 IF version) when analysed as a downconverter.

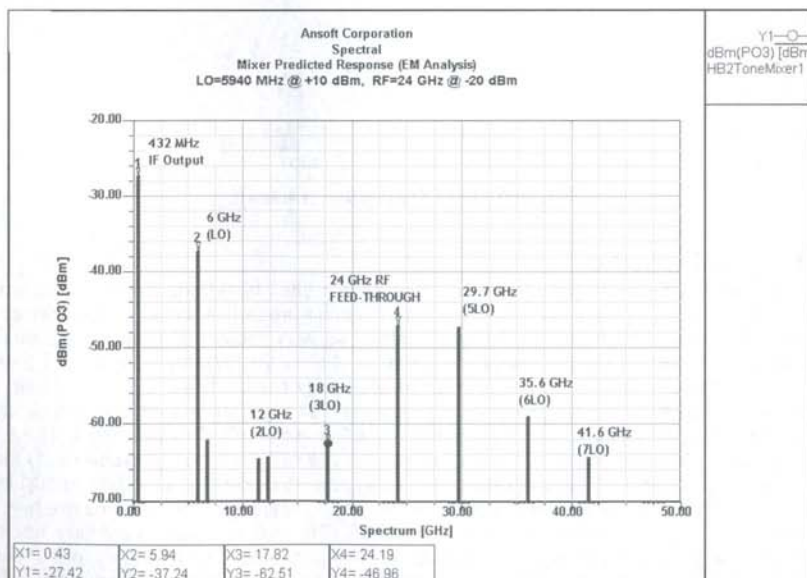


Figure 11: Predicted spectral response of the 24GHz harmonic mixer for a 432 IF

The above plot is for the mixer with a 24GHz input level of -20dBm, a 5940MHz LO at +10dBm and a 432MHz IF. Marker 1 is at the IF output, at a level of -27.4dBm, so the predicted conversion loss is 7.4 dB – a very respectable figure for a x4 harmonic mixer. The 6GHz LO feedthrough (Marker 2) is at -37dBm, which is 10dB below the desired IF signal level. Since the 6GHz input is at +10dBm, this shows an LO rejection of 47dB. The 2LO and 3LO responses are also very well attenuated. The 5LO, 6LO, and

7LO responses can also be seen, but are not of much concern as they are well attenuated and far removed from the desired IF. Marker 4 shows the 24 GHz signal (4LO) that appears at the IF output; at -47 dBm, this signal is suppressed by more than 25 dB.

The computer model of the 144MHz IF version of the mixer is nearly identical to that shown above.

A waveguide to microstrip transition

Since a waveguide filter is normally used on the RF port of the mixer, a WR42 (WG20) waveguide interface is desirable. This allows a waveguide filter to be bolted directly to the mixer without the need for a cable or WG-SMA transition. I decided to build a waveguide transition right into the mixer. A waveguide to coax transition is normally accomplished by soldering a small diameter pin to the microstrip, which in turn acts as a waveguide probe placed into the guide with a waveguide backshort roughly $\frac{1}{4}$ of a guide wavelength away from the probe. For a good discussion on this, see Paul Wade W1GHZ's QEX article [8].

After making a number of 24GHz WG to coax transitions, I quickly realized that the length of the probe can be very tricky to get just right! I was concerned that if a small wire was used as a probe, the preparation and placement of the probe might result in less than optimal performance (or too much tuning!). For this reason I chose a different approach: a direct microstrip to waveguide transition was designed where the microstrip itself becomes the probe for the transition. The design is illustrated in figure 12 below. The left side represents the 50 Ω microstrip line while the rectangular block to the right represents a WR42 (WG20) waveguide; the microstrip is extended into the guide to form a probe, with a waveguide back short roughly $\frac{1}{4}$ wavelength below the probe. The microstrip probe is widened to improve the impedance match over a relatively broad frequency range centred around 24GHz. The PC board itself is inserted into the guide and the groundplane on the reverse side of the PC board is removed to allow the fields to pass through the dielectric material.

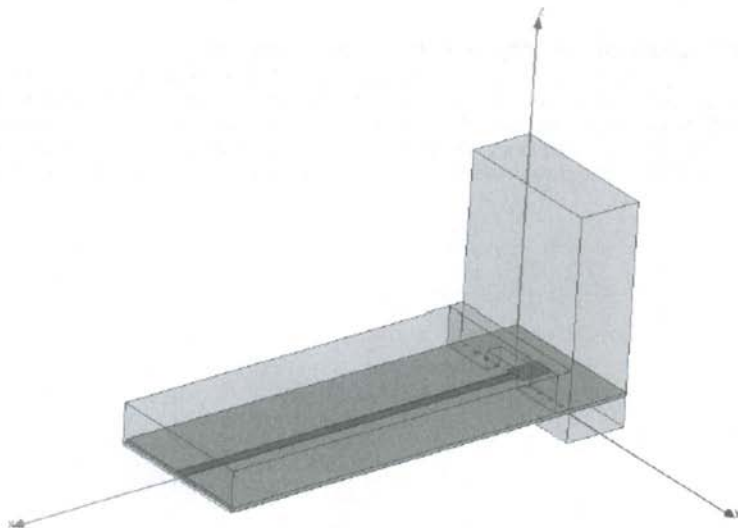


Figure 12: A microstrip to waveguide transition using microstrip probe

The above design was modeled using Ansoft HFSS. The model predicts less than a 0.3 dB insertion loss and a very good return loss for this design.

Assembling the mixer

A machined aluminum housing was designed to house the mixer assembly including the waveguide to microstrip transition. The WR42 (WG20) waveguide interface for the RF port incorporated directly into the mixer housing as described above, and the completed assembly is shown in figure 13. The diodes are not shown in this conceptual drawing but are placed near the centre portion of the microstrip transmission line. The LO input is the SMA connector at the lower edge and the connector at the right side is for the IF. The overall mixer measures less than 1.5 x 2 inches (38 x 50 mm). The printed circuit board is attached to

the metal base using a conductive epoxy. Following attachment of the board, the connectors are attached and the diodes are silver epoxied to the board.

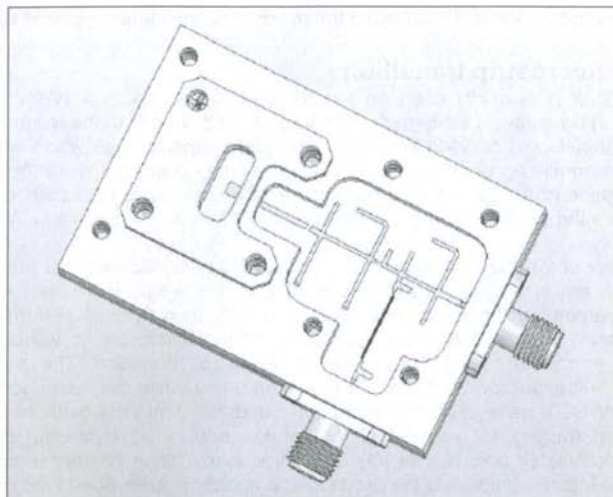


Figure 13: Mixer assembly

Measured results of prototype subharmonic mixer

The mixer was first measured as a downconverter with a 24.192GHz input of -20dBm and a $+10\text{dBm}$ LO. The plot in figure 14 shows a spectrum analyzer plot of the downconverted IF signal over a 50MHz span; the 144MHz IF signal measures -29.5dBm . This translates to a measured conversion loss of 9.5dB. When varying the RF input frequency, the mixer response is essentially flat to within 1 dB over an RF bandwidth of $\sim 500\text{MHz}$. The design works equally well at the 24.048GHz amateur band used in Europe and elsewhere.

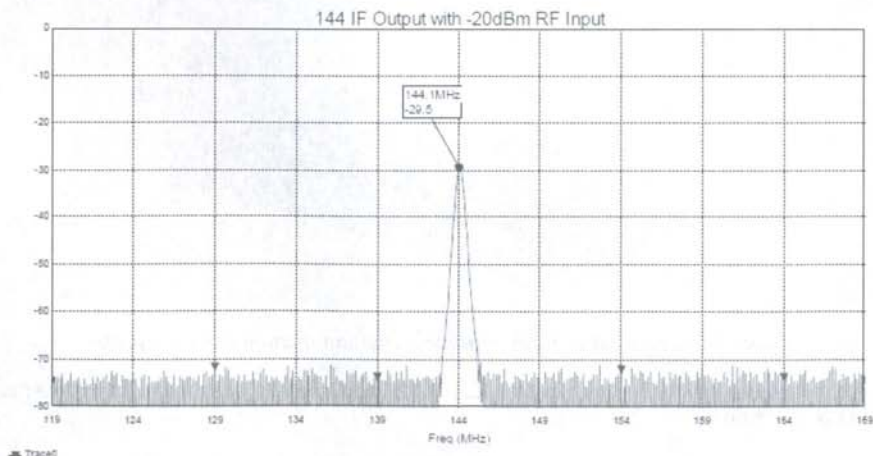


Figure 14: Prototype mixer: measured results as a receive mixer

Next, the prototype was measured as a transmit mixer with an output at 24.192GHz. The LO level is $+10\text{dBm}$ and the IF drive level was set at -10dBm . While slightly more output could be achieved by driving the mixer harder, this IF drive level represents the highest drive level without driving the mixer into compression. Note that as a transmit mixer, the conversion loss is again 9.5 dB. The measurement was made over a 100MHz output bandwidth, and no close-in spurious signals were seen.

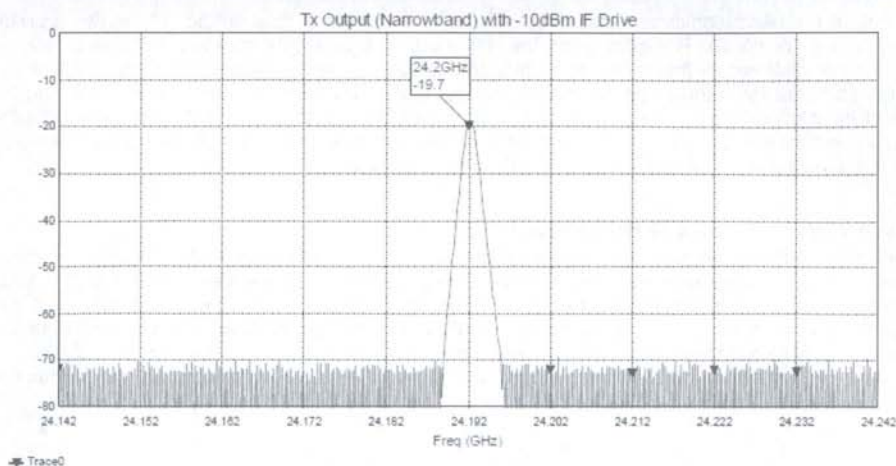


Figure 15: Transmit mode results for prototype mixer (narrowband)

Finally, the transmit bandwidth was widened in order to measure the rejection of the close-in 4xLO signal; results are shown in figure 16.

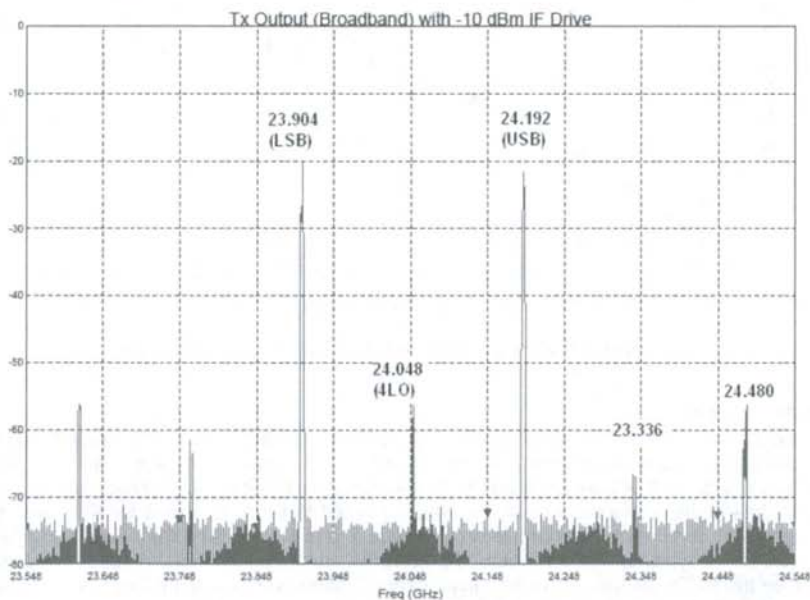


Figure 16: Transmit mode results for prototype mixer (broadband)

The above broadband spectrum analyzer measurement was made over a 1GHz bandwidth, and there are a number of undesired mixer signals, as expected. The desired RF signal at 24192MHz is the USB mixing product and is seen at 1½ divisions to the right of centre. The undesired LSB signal at 23.904GHz is equal in amplitude to the desired RF signal. As expected, there are also several additional mixing products at multiples of 144MHz away from our desired RF signal; however they are more than 35 dB below the desired. A surprising result can be viewed in this data. **Note that the 4LO signal (at the centre) is more than 35 dB below the desired 24.192GHz signal.** In most fundamental mode mixers the LO signal is at the same level as the desired USB RF signal. While RF filtering is still obviously required, the x4 subharmonic mixer should require less filtering than a standard type mixer. Lastly, one

comment on the mixer conversion loss is in order. The computer simulation (fig. 11) shows a predicted conversion loss of ~ 8 dB. The conversion loss figures on the prototype measure ~ 9.5 dB so there is a difference noted. Some of this is due to etching tolerances, etc but there are two additional factors that were unaccounted for in the computer model: the loss in the WG transition and the loss of the additional length of input microstrip between the WG transition and the actual mixer input. When these additional losses are included, the revised prediction for the conversion loss is almost 9 dB, which then agrees very well with the measured conversion loss of 9.5dB in the prototype.

Conversion loss vs. LO drive level

Like all mixers, the conversion loss varies as a function of the LO drive level. In order to measure conversion loss, a downconversion measurement was made on a mixer designed for a 144 MHz IF. Although +10 dBm LO is a convenient LO drive level, figure 17 shows that the conversion loss may be improved if one uses an LO of approximately +13dBm. This is also the level at which the mixer will be least affected by measurement errors or over-optimism about the true LO drive power.

When used as a transmit mixer, it has been found that with a +10dBm LO, the mixer output compresses with IF drive levels of 0 dBm to +5 dBm.

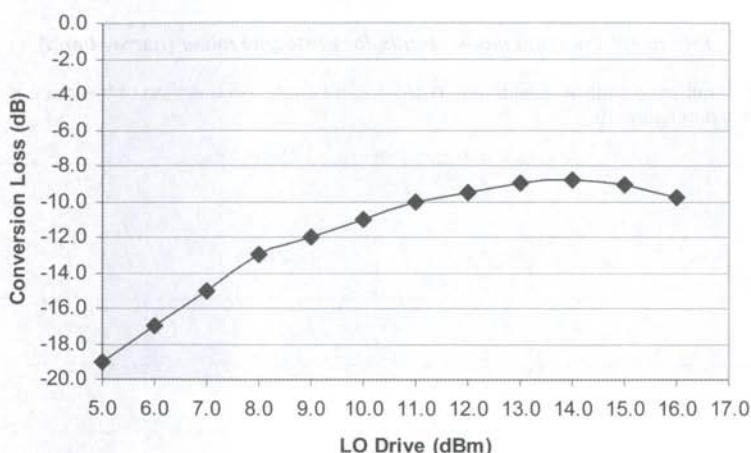


Figure 17: Measured conversion loss vs LO drive level

Additional efforts

The prototype subharmonic mixer was designed using Teflon-based Rogers 5880 board material. The fabrication costs using this material were a bit on the high side for amateur radio applications. In an effort to improve this, the mixer has been redesigned for the lower cost Rogers 4350 type material. While the 4350 material has higher dielectric losses, the tradeoff is felt to be a good one for the sake of a more affordable design. A number of mixers have now been built and tested using the new board material and the conversion loss is typically in the range of 10-12dB. All undesired signals are suppressed roughly as in the prototype mixer. The above conversion loss measurements were made on the newer version mixer. I am in the process of designing a similar subharmonic mixer for use at 47GHz.

Conclusion

A low loss x4 subharmonic mixer has been designed and demonstrated with favorable results. It's my hope that this design will make it easier for amateurs to get on the 24GHz amateur band. I've produced a limited quantity of mixers making them available to those who may be interested; please see www.w2ped.com for more information.

The author wishes to thank Al Ward W5LUA, Barry Malowanchuk VE4MA, and Brian Justin WA1ZMS for their thoughtful suggestions and inputs to the original paper presented at the 2008 Microwave Update conference. I also thank Ian White GM3SEK, for his helpful suggestions for this *DUBUS* article.

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Ein verlustarmer subharmonischer x4-Mischer für 24 GHz

von Paul Drexler, W2PED

Einleitung

Bei Mikrowellenfrequenzen wird die Erzeugung des LO-Signals typischerweise mit einem Quarzoszillator niedriger Frequenz oder einen Synthesizer gefolgt von einer Vervielfachung auf die endgültige LO-Frequenz vorgenommen. Billige Vervielfacher arbeiten gut bis ca. 10 GHz. Bei 24 GHz und darüber arbeiten viele der billigen MMIC-Bauteile einfach nicht gut. Einen LO im Grundmodus für 24 GHz aufzubauen ist etwas schwierig.

Eine andere Lösung (die hier beschrieben wird) besteht darin, einen subharmonischen Mischer zu verwenden. Dieser Mischertyp verwendet einen LO-Eingang, der niedriger ist, als die die normale LO-Eingangsfrequenz. Die Umsetzungsfunktion des Mixers vervielfacht effektiv die LO-Eingangsfrequenz mit dem Faktor n im Mischer, um die benötigte hohe Frequenz zu erreichen. Einen subharmonischen Mischer gibt es in Ausführungen mit einfacher LO-Hardware. Dies ist besonders hilfreich, wenn man auf ein höheres Frequenzband, wie 24 GHz, möchte. Hier wird ein subharmonischer Mischer höherer Ordnung ($n = 4$) für das 24-GHz-Amateurband vorgestellt, siehe Abb. 1). Der Umsetzungsverlust ist nur geringfügig höher als bei einem Mischer im Grundmodus und nur ein LO bei ca. 6 GHz ist nötig, um auf 24 GHz zu arbeiten. Verschiedene Frequenzmischungsschemata für den subharmonischen Mischer werden in Tabelle 2 für die Bänder 24048 MHz und 24192 MHz vorgeschlagen.

Abb. 1: Subharmonischer Mixer für 24 GHz von W2PED

HF Freq.	ZF Freq.	LO Eingang
24,048	144	5976
24,048	432	5904
24,192	144	6012
24,192	432	5940

Abb. 2: Verschiedene Frequenzmischungsschemata für 24 GHz

Eigenschaften von subharmonischen Mischern

Ein subharmonischer Mischer erlaubt uns, einen einfacheren LO zu verwenden, aber benötigt ein anspruchsvolleres Mischer-Design. Auch ist die Effektivität bei subharmonischen Mischern geringer, z.B. hat der subharmonische Mischer HP 11970A von Hewlett Packard einen Umsetzungsverlust von fast 20 dB! Zum Teil liegt das daran, dass dieser Mischer sehr breitbandig ist. Zum Glück wird im Amateurbereich meist schmalbandig gearbeitet und bei schmalbandigem Betrieb kann die Effektivität der Schaltung stark verbessert werden. Publizierte schmalbandige subharmonische $x2$ ($n = 2$) Mischer für 24 GHz haben einen typischen Umsetzungsverlust von 8 bis 10 dB, so dass wir einen höheren Verlust für einen subharmonischen $x4$ ($n = 4$) Mischer erwarten sollten. Zusätzlich zu dem höheren Umsetzungsverlust, kann ein subharmonischer Mischer aufgrund des internen Multiplikationsprozesses mehr unerwünschte Signale aufweisen. Bei breitbandigen kommerziellen Anwendungen kann dies ein Problem sein, nicht aber bei Amateurfunkanwendungen.

Subharmonische Mischer von anderen Autoren

Es ist wichtig darauf hinzuweisen, dass einige Amateure Arbeiten über subharmonische Mischer publiziert haben. Jim Dietrich, WA0RDX, hat ein neuartiges subharmonisches $x2$ Mischer-Design für 1296 MHz in der Oktober-Ausgabe 1978 des Ham Radio Magazins veröffentlicht [1]. Sein Mischer verwendet zwei Rücken-an-Rücken-Dioden in Serie mit einem resonanten Mikrostreifenleitungskreis. Der subharmonische Mischer für 24 GHz (und andere Bänder) von DB6NT ist ein weiteres Beispiel für einen subharmonischen $x2$ Mischer [2]. Brian Justin, WA1ZMS, hat einen subharmonischen Wellenleiter-Mischer bei seiner Arbeit auf 145 und 241 GHz eingesetzt [3] und diverse Mikrowellen-DX-Rekorde damit aufgestellt. Brian hat einen $x3$ Mischer als subharmonischen Mischer modifiziert, in dem er einen ZF-Tiefpassfilter hinzufügte und die Dioden vorspannte. Er hat auch eine schöne Diskussion verschiedener Dioden-Vervielfacher und Mischertypen im Tagungsband des Microwave Update 2005 veröffentlicht [4]. Diese Arbeit hier ist fokussiert auf ein 24-GHz-Mischer-Design, das Mikrostreifenleitungen verwendet. Andere auf Wellenleitern basierende Designs für subharmonische Mischer existieren und viele Artikel sind in DUBUS und anderen Publikationen veröffentlicht worden. Ich habe nicht intensiv nachgeforscht, aber einer dieser Artikel ist von Peter Riml, OE9PMJ, und beschreibt einen auf Wellenleitern basierenden subharmonischen Mischer für 47 GHz [5].

Ein Mischer höherer Ordnung

Ich war der Meinung, dass es nützlich sein könnte, einen auf Mikrostreifenleitungen basierenden subharmonischen Mischer für 24 GHz zu entwickeln, der ein $x4$ LO-Schema verwenden sollte (entsprechend obigem, $n = 4$). Ich dachte, dass, sofern erfolgreich, dieser den Aufbau eines 24-GHz-Transvertes etwas vereinfachen könnte, weil nur ein einziger LO für ca. 6 GHz für die typische 144 MHz ZF benötigt werden würde. Ich glaube, dass es einfacher ist, ein 6-GHz-LO-Signal zu erzeugen, als einen LO-Signal für 12 oder 24 GHz zu generieren. Das Schema hat also den Vorteil, einen niederfrequenten LO zu verwenden, aber der Nachteil ist, dass es ein komplexeres Design ist und dieses einen größeren Umsetzungsverlust aufweist, als ein Mischer im Grundmodus. Die am vielversprechendste Mischer-Topologie, die ich für ein subharmonisches $x4$ Mischer-Design gefunden habe, siehe Abb. 3, basiert auf der Arbeit, die in der IEEE-Literatur von A. Madjar beschrieben wurde [6].

Abb. 3: Subharmonischer Mischer - Blockschaltbild

Es gibt verschiedene wichtige Dinge, die bewirken, dass diese Mischerschaltung als effektiver subharmonischer Mischer arbeitet. Erstens arbeiten zwei Dioden „Rücken-an-Rücken“: die sogenannte „antiparallele“ Diode. Die antiparallele Anordnung unterdrückt zirkulierende Ströme auf der LO-Eingangsfrequenz und ihren geraden Harmonischen ($n=2, 4$ usw.), während gleichzeitig das Mischen mit diesen geraden Harmonischen gefördert wird. Das ermöglicht, dass ein gut designter Mischer mit niedrigem Umsetzungsverlust arbeiten kann, während gleichzeitig eine gute Unterdrückung einiger höherer LO-Harmonischer ermöglicht wird. Beide diese Eigenschaften stehen im Widerspruch zu dem, was man als Funkamateure normalerweise erwarten würde. Zweitens werden Filter in den HF-, ZF- und LO-Ports des Mixers eingesetzt. Jedes Filter ist so dimensioniert, dass die erwünschten Signale (d.h. HF, LO oder ZF) durchgelassen werden, während andere Signale, die intern im Mischer entstehen, reflektiert werden. Diese Filter werden im Detail weiter unten besprochen.

Wie bei den meisten HF-Schaltungen werden die besten Ergebnisse erhalten, wenn alle parasitären Induktivitäten und Kapazitäten so klein wie möglich gehalten werden. Bei 24 GHz bedeutet dies, *sehr*

kleine Komponenten zu verwenden. Bei einem Mischer ist es auch wichtig Dioden zu verwenden, die in ihren DC-Charakteristiken gut zueinander passen. Zum Glück haben einige kommerzielle Halbleiter-Hersteller gematchte Paare billiger Schottky-Dioden für hohe Frequenzen im Programm. Eine solches antiparalleles Diodenpaar ist die M/A COM MA4E1318, die in Abb. 4 gezeigt ist. Man beachte, dass die parasitäre Kapazität für dieses Diodenpaar unter 0,06 pF liegt!

Erfreulich ist, dass Dioden mit guter Leistung auf hohen Frequenzen auch in kleinen surface-mount Gehäusen angeboten werden, so dass kein Wire-Bonding nötig ist. Ein leitender Epoxykleber kann verwendet werden, um diese Dioden auf der PCB des Mischers zu befestigen. Unerfreulich ist, dass diese Bauteile dennoch nur 0,66mm lang sind, was wirklich klein ist! Siehe Abb. 4b. Ein gutes Mikroskop und eine ruhige Hand sind nötig, um mit ihnen arbeiten zu können.

Abb. 4a: MA/COM-Spezifikationen für einfache und antiparallele Mischerdiode

Abb. 4b: Antiparallele Mischerdiode MA/COM

Grundlagen der Arbeitsweise des Mischers

Der Mischer besteht aus den antiparallelen Dioden, dem HF-Filter, LO-Filter und ZF-Filter-Netzwerk, siehe Blockschaltbild in Abb. 3. Die Arbeitsweise des Mischers kann entweder als Empfangs- oder als Sendemischer beschrieben werden. Wir beziehen uns auf den Empfangsmischer, um die Dinge einfach zu halten. Es ist auch weniger verwirrend, wenn man die verschiedenen Bezugsfrequenzen als ungefähre Werte angibt. Beginnend mit dem LO-Eingang wird also ein externes 6-GHz-LO-Signal (ca. 10 dBm) bei der LO-Eingangsbuchse eingespeist. Ein schmalbandiges gekoppeltes Bandpassfilter ist auf 6 GHz zentriert. Es filtert die LO-Energie, aber noch wichtiger, es reflektiert alle anderen Frequenzen. Wenn die 6-GHz-LO-Energie das Diodenpaar trifft, werden intern LO-Harmonische generiert. Die x4-6-GHz-Energie wird unsere erwünschte Hochfrequenz-LO im Mischer. Weitere erzeugte LO-Harmonische sind die x2-, x3-, x5- und höheren Vielfachen. Alle diese Signale werden im Wesentlichen zum Mischer zurückreflektiert und dies verbessert nach entsprechender Optimierung die Mischer-Effektivität, was essentiell für einen niedrigen Umsetzungsverlust ist. Das HF-Netzwerk hat zwei Aufgaben: Erstens, es arbeitet als verlustarmes breitbandiges Bandpass-Filter, zentriert auf 24 GHz. Das Bandpassfilter liefert bis zu einem gewissen Grad eine Außerbandunterdrückung, aber noch wichtiger, es hält die verschiedenen LO-Signale von der Abstrahlung über den HF-Eingang des Mischers ab.

Es wirkt auch reflektierend auf die LO-Energie, die zurück zu den Dioden reflektiert wird. Zum Schluß lässt die ZF-Sektion des Mischers die ZF-Ausgangssignale des Mischers durch, und genau wie in den anderen Abschnitten des Mischers, reflektiert sie die LO-Energie zurück auf die Mischerdioden. Dies wird mittels Leerlauf-Kreisen mit hohem Q (offene Mikrostrip-Stubs) realisiert. Jeder ist auf der benötigten Frequenz resonant und reflektiert Energie zurück auf die Dioden. Wenn man auf den endgültigen Aufbau des Mischers blickt, kann man erkennen, dass die ZF-Sektion doch nicht ganz so simpel ist. Jede der Haupt-Sektionen des Mischers wird im Folgenden detailliert beschrieben.

HF-Filter

Das HF-Bandpassfilter beginnt zuerst als einfache Anordnung aus zwei kurzgeschlossenen $\lambda/4$ -Leitungen, die Resonatoren auf 24 GHz bilden. Die Resonatoren werden direkt durch eine weitere Leitung gekoppelt, die ca. $\lambda/4$ lang ist. Alle Leitungen haben ca. 50 Ω . Die Schönheit dieser Filter-Topologie besteht darin, dass keine eng gekoppelten Leitungen nötig sind, wie es bei randgekoppelten Filtern der Fall wäre, so dass das physikalische Layout recht tolerant ist. Die resultierende Kurve ergibt einen breitbandigen Bandpass bei 24 GHz.

Abb. 5: Einfaches HF-Filter

Abb. 6: HF-Eingangsfiler mit Stubs für die LO-Harmonischen

Als nächstes werden dem Filter resonante Stubs am Eingang und Ausgang zugefügt, um eine Kurzschlusscharakteristik auf der LO- und 2xLO-Frequenz zu erhalten. Jeder der offenkreisigen Mikrostreifenleitungs-Stubs hat ca. $\lambda/4$ -Länge auf der Designfrequenz. Diese wirken wie Fallen-Filter auf der LO- und 2xLO-Frequenz. Zuerst wurde das HF-Filter mittels ungenauer Schätzung per Handrechnung abgeschätzt und dann mit der Studentenversion der Ansoft Designer Software [7] optimiert. Die Längen der Leitungen wurden für beste Anpassung und geringste Dämpfung auf 24 GHz justiert und auf beste Resonanz für jede Falle. Die simulierte Filterkurve ist in Abb. 7 gezeigt. Die Y1-(S21)-Kurve zeigt, dass eine geringe Einfügedämpfung von 0,25 dB für 24 GHz vorhergesagt wird, wobei die LO- und 2xLO-

Frequenzen stark unterdrückt (reflektiert) werden. Die Y2-(S11)-Spur ohne Marker zeigt den Returnloss für jede Frequenz. Das HF-Filter ist für einen sehr guten Returnloss auf 24 GHz designt und eine reflektierende Impedanz auf der LO- und 2xLO-Frequenz. Die 3 λ /4-Resonanz der LO-Stubs liefert eine zusätzliche Nullstelle bei der 3LO-Frequenz (ungefähr 18 GHz), so dass wir für zwei Stubs 3 Nullstellen für den 6-GHz-LO und seine stärkeren Harmonischen niedriger Ordnung erhalten.

Abb. 7: Simuliertes HF-Filter mit Stubs

LO-Filter

Das LO-Bandpass-Filter ist ein einfaches randgekoppeltes Design aus zwei λ /4-Leitungen auf der 6-GHz-LO-Frequenz. Ein komplexeres Filter ist nicht nötig, weil das LO-Filter einfach nur die LO durchlassen muß und auf allen anderen Frequenzen reflektiv wirken soll. Die Mittenfrequenz wurde für 6 GHz designt und die Bandbreite wurde weit genug gemacht, um andere erwünschte LO-Frequenzen in diesem Bereich abzudecken. Die größere Bandbreite stellt auch sicher, dass Abweichungen, die beim Ätzprozess der Platine auftreten können, die Leistung nicht mindern werden. Der gekoppelte Anteil der Leitung des Filters ist ca. 10 mm (0.4 Zoll) lang.

Um eine 50 Ω -Impedanzanpassung am Ein- und Ausgang des Filters zu haben, müssen die gekoppelten Transmissionsleitungen eine höhere Impedanz aufweisen und deshalb sehr schmal sein. Die Leitungen sind idealerweise 4 mils (0.1 mm) breit und haben 4 mils (0.1 mm) Abstand, wenn auf einem 10 mil Board aus Teflonsubstrat gearbeitet wird. Aber diese Linienstärken und Abstände sind nur schwer reproduzierbar zu ätzen, selbst von einem kommerziellen Hersteller. Stattdessen habe ich mich also lieber für einen kleinen Kompromiss beim Design entschieden und die gekoppelte Sektion der Linien breiter und etwas weiter auseinander liegend gemacht, um sie besser herstellbar zu machen. Das verschlechtert die Anpassung, aber die beiden Stubs am Eingang und Ausgang korrigieren das Problem.

Abb. 8: Layout des gekoppelten LO-Filters

Das Designen von randgekoppelten Filtern kann sehr schwierig sein, wenn das Design nicht alle Kopplungseffekte berücksichtigt. Am besten man macht dies in einem elektromagnetischen (EM) Simulator. Eine EM-Simulation zeigte, dass das erste Design in der Frequenz versetzt lag und eine Einfügedämpfung von ca. 3 dB haben würde. Die erwünschte Mittenfrequenz wurde durch Justierung der Filterlänge und erneutes Analysieren im EM-Simulator erreicht. Die berechnete Einfügedämpfung des LO-Filters beträgt nun weniger als 0.5 dB bei 6 GHz.

ZF-Netzwerk

Die ZF-Sektion des Mischers könnte man für den „NF-Teil“ des Designs halten, aber tatsächlich ist sie etwas mehr als das. Die ZF-Sektion speist die LO-Energie in die Mischer-Dioden, leitet die ZF-Signale zu/von den Dioden und reflektiert die HF-, LO- und 3xLO-Energie zurück zu den Dioden. Die HF-Energie und alle LO-Produkte werden aus zwei Gründen auf die Dioden zurückreflektiert:

- um unerwünschte Signale vom ZF-Ausgang abzublocken (d.h. HF und LO-Signale abzuschwächen),
- die Reflektion aller Signale (außer der ZF) zurück auf die Dioden ergibt weniger Umsetzungsverlust.

Jeder der offenkreisigen Mikrostreifenleitungsstubs ist ca. λ /4 auf der Designfrequenz und kann auch als Leerlaufkreis mit hohem Q angesehen werden. Das ist genau wie bei den resonanten Stubs, die in der Vergangenheit in den 432->1296-MHz-Varaktor-Verdreifacher-Kreisen eingesetzt wurden. Die Höhe der Mischerleistung ist extrem abhängig von den Längen der Stubs. Sollte ein Stub in der Länge um mehr als ein paar mils (Bruchteile eines Millimeters) abweichen, wird sich die Leistung des Mischers schnell verschlechtern. Die höherfrequenten Stubs liegen am dichtesten an den Dioden, um die Strecke der reflektierten Transmission von HF- und 3xLO-Energie kurz zu halten. Alle Längen der verbindenden Leitungen wurden auf bestmögliche Leistung optimiert.

Abb. 9: Netzwerk für die ZF/LO-Einspeisung

Zusammenschaltung

Der Mischer wurde zuerst für 10 mil (0.25mm) dickes HF-Board-Material 5880 von Rogers entwickelt. Das Material wurde wegen der sehr guten HF-Eigenschaften, inkl. niedriger dielektrischer Verluste (Dielektrizitätsverlust ~0.0009 auf 10 GHz) ausgewählt. Ein 10 mil dickes Material wurde gewählt, um die Breiten der Transmissions-Leitungen angemessen für die Arbeitsfrequenz zu behalten. Wenn stattdessen 5 mil Material verwendet würde, müsste das LO-Bandpassfilter extrem dünne Bahnen haben. Sobald das

Grunddesign einer jeden Sektion fertig war, bestand der nächste Schritt darin, alles zusammen zu verbinden und die Schaltung mit Hilfe des Ansoft Designers, einem nichtlinearen Schaltungssimulator, zu analysieren. Ein LO-Pegel von +10dBm wurde verwendet. Die ersten Ergebnisse waren nicht gerade gut... der vorhergesagte Einfügeverlust betrug über 20 dB! Das lag an 1.) Wechselwirkungen zwischen den Sektionen, 2.) jedes der Filter-Netzwerke war für eine ideale 50 Ω -Last entwickelt worden, die Dioden aber repräsentieren eine davon abweichende Impedanz und 3.) da jeder der Stubs hohe Güte (Q) hat, war eine weitere Optimierung unbedingt nötig. Der nächste Schritt war also die Optimierung eines jeden Stubs und der Länge der Transmissionsleitung, um beste Leistung zu erhalten. Das Ziel war, den Mischerumsetzungsverlust so gering wie möglich zu machen, während die unerwünschten Signale so niedrig wie möglich gehalten werden sollten. Zum Glück verbessert sich generell die Mischereffektivität, sobald eines der unerwünschten Output-Signale vermindert wird. Sobald das erste Schaltungsmodell zufrieden stellend war, könnte dann eine genauere EM-Analyse der physikalischen Struktur gemacht werden. Wie zuvor erwähnt, erbringt die elektromagnetische Analyse (EM-Analyse) genauere Ergebnisse, da sie elektromagnetische Effekte berücksichtigt, inkl. der gegenseitigen Kopplung zwischen den Leitungen. Der Nachteil ist, dass dies eine echte Rechenleistung benötigende Aufgabe für den PC darstellt, die sehr zeitaufwendig wird.

Ein weiteres Problem ist, dass es nicht einfach möglich ist, EM-Analyse und Optimierung gleichzeitig zu machen, so dass dies ein mehrstufiger Prozess wird. Ich entschied mich dafür, dass es die Mühe wert ist, diesen extra Schritt zu gehen und das Design feinabzustimmen mittels der genaueren EM-Analyse. Das würde die Wahrscheinlichkeit für einen sofortigen Erfolg beim Prototyp in der Praxis erhöhen. Unter Verwendung dieser präziseren Methoden wurden zwei Mischer entwickelt, einer für 144 MHz ZF und ein weiterer für 432 MHz ZF. Abb 10 zeigt das Layout des Mischers. Der HF-Eingang ist links, der LO-Eingang unten und ZF rechts. Die Länge des Mischers ist gerade mal gut 0.75 Zoll (1.9cm). Das LO-Filter scheint eine einzige dicke Linie zu sein, sofern man das Bild nicht vergrößert anschaut.

Abb. 10: Prototyp des Mischerlayouts

Der Plot in Abb. 11 zeigt das vorhergesagte Spektrum des Mischers (in der 432-MHz-ZF-Version), wenn er als Down-Konverter analysiert wird.

Abb. 11: Simuliertes Spektrum des 432-MHz-ZF-Mischers

Dieser Plot ist für einen Mischer, der mit einem Eingangspegel von -20 dBm auf 24 GHz und einem LO auf 5940 MHz mit +10 dBm und 432 MHz ZF analysiert wurde. Marker 1 ist beim ZF-Ausgang und liegt bei einem Pegel von -27.4 dBm, so dass der vorhergesagte Umsetzungsverlust bei 7.4 dB liegt, was ein sehr respektables Ergebnis für einen harmonischen $\times 4$ -Mischer ist. Der 6-GHz-LO-Durchschlag (Marker 2) liegt bei -37 dBm, was 10 dB unter dem erwünschten ZF-Signalpegel ist. Da der 6-GHz-Input bei +10 dBm liegt, ergibt dies eine LO-Unterdrückung von 47 dB. Die 2LO- und 3LO-Signale sind ebenfalls sehr gut unterdrückt. Die 5LO-, 6LO- und 7LO-Signale sind ebenfalls sichtbar, aber nicht beunruhigend, da sie alle sehr gut im Vergleich zum erwünschten ZF-Signal unterdrückt sind und weit entfernt liegen. Marker 4 zeigt das 24-GHz-Signal (4LO), das am ZF-Ausgang erscheint. Bei -47 dBm ist dieses Signal um mehr als 25 dB unterdrückt. Das Computermodell der 144-MHz-ZF-Version des Mischers ist nahezu identisch mit dem hier gezeigten.

Ein Übergang von Wellenleiter auf Mikrostreifenleitung

Da ein Wellenleiter-Filter normalerweise am HF-Port des Mischers eingesetzt wird, wäre ein WR42 (WG20)-Übergang wünschenswert. Das würde ein direktes Anflanschen eines Wellenleiter-Filters an den Mischer ohne ein Kabel oder einen SMA-Wellenleiter-Adapter ermöglichen. Ich entschied mich dafür, einen Wellenleiterübergang direkt in den Mischer einzubauen. Ein Übergang von Wellenleiter auf koaxial wird normalerweise durch Auflöten eines kleinen Stifts (Pin) auf die Streifenleitung hergestellt, der als Koppelsonde im Wellenleiter fungiert, der wiederum ungefähr im Abstand von $\lambda/4$ von der Sonde kurzgeschlossen ist. W1GHZ erläutert diese Anordnung gut in einem Artikel in QEX [8].

Nachdem ich eine Anzahl von Übergängen von WL auf Koax für 24 GHz angefertigte hatte, habe ich bemerkt, dass es sehr schwierig ist, die richtige Länge der Sonde hinzubekommen. Ich war besorgt, dass wenn ein kleiner Draht als Sonde benutzt würde, die Anfertigung und Platzierung der Drahtsonde in suboptimaler Leistung resultieren könnte (oder zuviel Abstimmerei!). Deshalb habe ich einen anderen Ansatz für dieses Problem gewählt. Es wurde ein direkter Übergang von Mikrostreifenleitung auf Wellenleiter entwickelt, bei der die Streifenleitung selbst die Sonde für den Übergang bildet. Die

Ausführung ist in Abb. 12 gezeigt. Die linke Seite ist die 50 Ω -Streifenleitung, während der rechteckige Block rechts ein WR42 (WG20) Wellenleiter ist. Die Streifenleitungs-sonde ragt in den Wellenleiter hinein, ein Wellenleiterkurzschluß befindet sich ungefähr in $\lambda/4$ Entfernung unterhalb der Sonde. Die Mikro-streifenleitersonde ist aufweitert, um die Impedanzanpassung über einen relativ breiten Frequenzbereich um das Zentrum von 24 GHz zu verbessern. Die Platine selbst wird in den Wellenleiter eingeführt. Die Massefläche auf der Unterseite der Platine wurde entfernt, damit die Felder durch das dielektrische Board dringen können.

Abb. 12: Ein Übergang von Mikrostriifenleitung auf Wellenleiter mit Mikrostriifenleitungs-sonde

Das obige Design wurde mit der Software Ansoft HFSS modelliert. Das Modell sagt weniger als 0.3 dB Einfügedämpfung und einen sehr guten Returnloss für dieses Design voraus.

Zusammenbau des Mischers

Ein gefrästes Aluminium-Gehäuse wurde entwickelt, um den Mischer und den Übergang von Wellenleiter auf Mikrostriifenleitung aufzunehmen. Der WR42-(WG20)-Wellenleiterübergang für den HF-Port wird direkt in das Mischergehäuse, wie oben beschrieben, eingebaut. Der komplette Aufbau ist in Abb. 13 gezeigt. Die Dioden sind in dieser Konzeptzeichnung nicht gezeigt, sie befinden sich in etwa der Mitte der Mikrostriifenleitung. Der LO-Eingang ist die SMA-Buchse unten links, die Buchse rechts ist für die ZF. Die Maße über alles betragen weniger als 1.5 x 2 Zoll (38 x 50mm). Die PCB wird mittels leitfähigem Epoxidkleber auf den Metallboden des Gehäuses geklebt. Danach werden die Buchsen eingebaut und die Dioden mit Silberleitkleber aufs Board geklebt.

Abb. 13: Aufbau des Mischers

Meßergebnisse des Prototyps des subharmonischen Mischers

Der Mischer wurde zuerst als Down-Konverter mit einem LO-Input von -20 dBm auf 24.192 GHz und einem +10 dBm LO gemessen. Der Plot in Abb. 14 zeigt das Bild des herunterkonvertierten ZF-Signals über eine Breite von 50 MHz auf einem Spektrumanalyzer. Das 144-MHz-ZF-Signal liegt bei -29.5 dBm, was einem Umwandlungsverlust von 9.5 dB entspricht. Die Änderung der Eingangsfrequenz über eine Bandbreite von 500 MHz zeigt, dass die Mischer-Kurve durchweg flach ist, d.h. innerhalb von 1 dB. Das Design arbeitet gleich gut auf dem europäischen 24.048-GHz-Band.

Abb. 14: Meßergebnisse des Mischer-Prototyps als Empfangsmischer

Als nächstes wurde der Prototyp als Sendemischer mit einem Output auf 24.192 GHz gemessen. Der LO-Pegel beträgt +10 dBm und der ZF-Treiberpegel wurde auf -10 dBm eingestellt. Obwohl ein wenig mehr Ausgangsleistung durch mehr Ansteuerung erreicht werden kann, stellt dieser ZF-Pegel den höchsten Pegel dar, bei dem der Mischer noch nicht in die Kompression gerät. Man beachte, dass beim TX-Mischer der Umwandlungsverlust auch wieder 9.5 dB beträgt. Die Messung wurde über eine Bandbreite von 100 MHz vorgenommen, im Nahbereich konnten keine unerwünschten Signale gesehen werden, siehe Abb. 15.

Abb. 15: Meßergebnisse des Mischer-Prototyps beim Senden (schmalbandig)

Zum Schluß wurde die TX-Bandbreite erhöht, um die Unterdrückung des 4xLO-Signals zu messen, die Ergebnisse zeigt Abb. 16.

Abb. 16: Meßergebnisse des Mischer-Prototyps beim Senden (breitbandig)

Die breitbandige Messung wurde auf einem Spektrumanalyzer über 1 GHz Bandbreite durchgeführt. Dort sind erwartungsgemäß einige unerwünschte Mischsignale zu sehen. Das erwünschte HF-Signal bei 24192 MHz ist das USB-Mischprodukt und ist bei $1\frac{1}{2}$ Skalenteilen rechts der Mitte zu sehen. Das unerwünschte LSB-Signal bei 23.904 GHz ist gleich groß in der Amplitude wie das erwünschte Signal. Es gibt diverse weitere Mischprodukte bei Vielfachen von 144 MHz oberhalb unseres erwünschten HF-Signals. Sie liegen aber mehr als 35 dB unterhalb des erwünschten Signals. Ein überraschendes Ergebnis kann man aus diesen Daten ersehen: **Man beachte, dass das 4LO-Signal (in der Mitte) mehr**

als 35 dB unterhalb des erwünschten 24.192-GHz-Signals liegt. Bei den meisten Mischern im Grundmodus ist das LO-Signal auf dem gleichen Niveau wie das erwünschte USB-HF-Signal. Zwar ist eine HF-Filterung offensichtlich immer noch nötig, aber ein subharmonischer x4-Mischer sollte doch weniger Filteraufwand benötigen als ein Standard-Mischer. Nun noch ein Kommentar zum Mischerumwandlungsverlust. Die Computersimulation in Abb. 11 zeigt einen vorhergesagten Verlust von ca. 8 dB. Die Messwerte für den Prototypen liegen bei 9.5 dB, es gibt also eine Differenz. Eine Ursache sind Toleranzen beim Ätzen etc., aber es gibt noch zwei weitere Faktoren, die im Computermodell unberücksichtigt blieben: Der Verlust des Wellenleiterüberganges und der Verlust der zusätzlichen Länge der Eingangstreifenleitung zwischen Wellenleiterübergang und eigentlichem Mischereingang. Wenn diese zusätzlichen Verluste berücksichtigt werden, liegt der nun vorhergesagte Umwandlungsverlust bei ganz knapp unter 9 dB, was dann sehr gut mit dem beim Prototyp gemessenen Umwandlungsverlust von 9.5 dB übereinstimmt.

Umsetzungsverlust in Abhängigkeit vom LO-Pegel

Wie bei allen Mischern ist auch hier der Umsetzungsverlust abhängig vom LO-Pegel. Um den Umsetzungsverlust zu bestimmen, wurde eine Heruntermischungsmessung mit einem Mischer für 144 MHz ZF gemacht. Obwohl +10 dBm ein geeigneter LO-Pegel ist, zeigen aber die Daten in Abb. 17, dass der Umsetzungsverlust verbessert werden kann, wenn man einen LO mit ca. +13 dBm verwendet. Dies ist auch der Pegel, bei dem der Mischer am wenigsten durch Messfehler oder Überschätzung der wirklichen LO-Treiber-Leistung beeinflusst wird. Bei Verwendung als TX-Mischer wurde herausgefunden, dass das Ausgangssignal des Mixers mit einem +10 dBm LO in die Kompression gerät, wenn die ZF-Treiberpegel zwischen 0 und +5 dBm liegen.

Abb. 17: Gemessener Umsetzungsverlust vs. LO-Treiberpegel

Weiteres

Der Prototyp des subharmonischen Mixers wurde mit auf Teflon basierendem Platinenmaterial 5880 von Rogers entwickelt. Die Kosten für dieses Material sind etwas hemmend für Amateurfunkapplikationen. Um dies zu verbessern, wurde der Mischer für das billigere Material 4350 von Rogers nachentwickelt. Obwohl das 4350-Material höhere dielektrische Verluste hat, ist es doch ein guter Kompromiss, da das Design nun erschwinglicher ist. Eine Anzahl von Mischern wurde mit dem neuen Board-Material gebaut und getestet. Der Umwandlungsverlust beträgt typisch zwischen 10 und 12 dB. Alle unerwünschten Signale sind ungefähr so unterdrückt wie beim obigen Prototyp. Die obigen Messungen für den Umsetzungsverlust wurden mit der neueren Mischer-Version vorgenommen.

Ich bin inzwischen dabei, einen ähnlichen subharmonischen Mischer für die Verwendung auf 47 GHz zu entwerfen und werde im Erfolgsfall darüber hier berichten.

Zusammenfassung

Ein verlustarmer subharmonischer x4 Mischer wurde entwickelt und mit guten Ergebnissen gemessen. Ich hoffe, dass dieses Design den Zugang für Amateure zum 24-GHz-Band erleichtert. Ich habe eine Anzahl von Mischern aufgebaut, die an interessierte Amateure abgegeben werden können. Weitere Infos gibt es unter www.w2ped.com.

Der Autor möchte Al Ward, W5LUA, Barry Malowanchuk, VE4MA, und Brian Justin, WA1ZMS, für ihre Vorschläge und Anregungen für den Original-Artikel, der auf der MUD 2008 in den USA präsentiert wurde, sowie Ian White, GM3SEK, für seine hilfreichen Vorschläge bei diesem *DUBUS*-Artikel.

Referenzen

Siehe engl. Text oben.

Herstellung von HL-Flanschen und deren Montage an Hohlleiter

von Philipp Prinz, DL2AM

Zu meinen Baubeschreibungen über die Transverter mit Wendeverstärker [1], gab es viele Gespräche und das Problem war des Öfteren, wie man wohl die dafür nötigen HL mit Flanschen anfertigen könnte. Professionelle Flansche mit einer Rechteck-Ausstanzung stehen den OMs nämlich meistens nicht zur Verfügung. Von diesen Transvertern mit Wendeverstärker habe ich inzwischen bereits 5 Stück gefertigt und einige Erfahrungen damit gesammelt (Abb. 1 + 2).

Auch in Tschechien, Tri Studne (Dreibrunnen) wo ich schon dreimal bei einem GHz-Treffen war, hielt ich beim letzten Besuch einen Vortrag über einen 76 GHz-Transverter mit Wendeverstärker. Wobei Milan, OK2IMH, bei der Übersetzung sehr hilfreich war.

Vor ein paar Wochen habe ich die Webseite von quinstar [2] angeschaut und bekam dabei feuchte Augen. Auf dieser Webseite können alle Bausteine, Geräte sowie alle Werte und Parameter von diesen eingesehen werden, die wir für unser außergewöhnliches Hobby benötigen. Erstaunt war ich unter anderem, dass es da einen Übergang gibt von Rechteck- auf Rund-HL, der nur 0,7 dB Loss bei 60 – 90 GHz hat.

Bei Conrad Elektronik [3] gibt es Messing-Wellen mit 20 mm Durchmesser. Mit diesen können, wenn eine Drehmaschine vorhanden ist, leicht HL-Flansche mit der Bezeichnung WG387 hergestellt werden. Flansche mit einem angedrehten Ansatz, siehe Abb. 2 a, eignen sich besonders gut. Von einem professionellen Flansch, wie zum Beispiel beim WG387, bei dem die Pass-Stifte herausgezogen wurden, können die 8 Bohrungen auf einen selbst gedrehten Flansch abgebohrt werden. Ich verwende nur noch Flansche mit 20 mm Durchmesser und einer Stärke von 4 mm, ähnlich wie der mit der Bezeichnung WG387, die 4 UNC 4-40 Gewinde in der vertikalen und horizontalen Achse für WR 19 bis WR 4 haben. Für die Fixierstifte, die auch bei Conrad als Stahldraht zu bekommen sind, haben diese Flansche noch zwei 1,5mm-Bohrungen und deren Gegenseite zwei 1,6 mm-Bohrungen in der 45°-Achse zur Ebene. Die Verbindungs-Schrauben dafür haben auch ein 4-40 Gewinde und der Innen-Sechskant hat eine Größe von 5/64 Zoll oder 3/32 Zoll. Diese sind nicht so leicht zu beschaffen.

Mit dieser Technik ist es möglich, bei Verwendung von professionellen Messgeräten, HL-Übergängen und Flansch-Adaptern von WG387 auf WG383 genauere Messergebnisse zu bekommen. Für die Flansche zur Spiegel-Montage an den Transverter sind bei mir vier 3-mm-Bohrungen in 45° zur Ebene vorgesehen, die auch bei den Procom-Spiegeln vorhanden sind. Der Transverter-Flansch hat vier M3-Gewinde in 45° zur Ebene. Dadurch habe ich ein System, in dem ich alle Spiegel tauschen kann. Wenn bei den Flanschen die zwei Fixier-Stifte mit 1,50 mm Durchmesser verwendet werden, müssen die dafür vorgesehenen Bohrungen sehr genau sein. In einer Zeichnung von DB6NT weist er schon 1992 auf diese Flansche hin. Das Anbringen dieser Flansche auf die HL WR 19, WR 12 und WR 4 ist gut möglich, wenn nachstehendes beachtet wird:

Beim WR 19 ist das Übereckmaß 7,75 mm, WR 12 5,95 mm, WR 8 4,15 mm und WR 4 3 mm. Wenn zum Beispiel beim WR 12 in den Flansch eine Mittelbohrung von 5,9 mm gebohrt wird (Abb. 3), auf der Innenseite mit einer konischen Reibahle (www.selva.de) die Kante leicht gebrochen wird, dann kann der HL auf den Flansch gedrückt werden bis er auf der Vorderseite bündig ist. Anstelle der konischen Reibahle kann auch ein Universal-Entgrater benutzt werden (Abb. 4). Vorher ist noch der HL an der Vorderseite rechtwinklig abzdrehen, fräsen oder feilen.

Beim Raufdrücken des HL auf den Flansch darf sich das Kupfer an den Kanten nicht zurückschieben, da sonst der Flansch nicht rechtwinklig zum HL steht und auch nicht fest sitzt, was aber sehr wichtig ist, um eine symmetrische Anordnung zu bekommen. Es ist aber auch möglich, beim HL die Kanten leicht konisch anzufeilen, um dann den Durchmesser der Bohrung für den Flansch festzustellen.

Wenn jetzt der HL mit dem Flansch in die Schieblehre gelegt wird (Abb. 5), sieht man, ob sich die Bohrungen in der Rotation an der richtigen Stelle befinden. Wenn notwendig, muss der Flansch noch nachgedreht werden. Jetzt sollte nochmals durch Rotation des HL der rechte Winkel des Flansches festgestellt werden (Abb. 6). Die Stirnseite des HL wird mit einem Folien-Filzschreiber kräftig bemalt, so dass beim Löten kein Zinn in das innere des HL laufen kann (Abb. 7 + 8). Nun den Flansch mit HL auf

eine ebene Platte stellen. Am besten verwendet man eine Keramik-Fliese dafür, um keinen Wärmeverlust zu haben, so dass das Zinn bis bündig Außenseite Flansch verlaufen kann. Mit einem Gasbrenner wird nun der Flansch mit HL erhitzt, bis das Zinn schmilzt und die gesamte Bohröffnung voll gelaufen ist (Abb. 9). Nun ist eine Seite geschaffen. Wenn auf der gegenüber liegenden Seite auch ein Flansch angebracht werden muss, ist die gleiche Prozedur nochmals zu machen. Bevor nun der zweite Flansch erhitzt wird, wird auf den oberen Flansch ein Stück Kupferplatte oder ähnliches gelegt, um ein Schmelzen des Zinns von dem oberen Flansch zu verhindern (Abb. 10).

Nun habe ich einen HL WR 12 mit ähnlichen Flanschen wie WG387. Ohne Dreh- oder Fräsmaschine ist es schwierig, die Flansche an der Außenseite plan zu bekommen. Ein geübter Feiler schafft es vielleicht (Abb. 11). Mit einem Skalpell wird nun vorsichtig die HL-Öffnung frei gelegt. Wenn von einem Rechteck-HL auf einen Rund-HL übergegangen wird, und umgekehrt, nehme ich einen Kegel-Fräser (Abb. 12) und fräse den Rechteck-HL soweit auf, bis gerade der innere Durchmesser des Rund-HL bündig ist. So können bei den HL einschließlich WR 19 und kleiner die Flansche befestigt werden. Größere HL als WR 19 mit Flanschen ähnlich WG387 können nach dieser Methode nicht mehr hergestellt werden.

Beim Reinigen von Hohlleitern sollte keine Watte, sondern ein Leinenstoff in die Drahtschlinge gelegt werden, der dann mit Aceton oder ähnlichem getränkt und durch den HL gezogen wird. Bei einer Watte bleiben immer Fussel zurück, die beim Beschlagen des HL Dämpfungen verursachen können (Abb 13).

Ergänzungen

Bei Leistungs-Messungen mit einem Thermo-Messkopf ist es vorteilhaft, einen Isolator mit bekanntem Loss (Dämpfung) zu benutzen, da dieser eine Bandpass-Charakteristik hat und Oberwellen nicht so stark in das Messergebnis einfließen lässt. Beim Leistungsmessen konnte ich diese Bandpass-Charakteristik eindeutig feststellen (Abb 14).

Auch habe ich 3 Harmonic-Mixer und einen subharmonischen Mixer für 122 GHz aufgebaut, wie von mir schon im Dubus 4/2006 unter „Neues Konzept“ [4] beschrieben. Erstere sind mit einer Single-Zero-Bias-Diode HSCH 9161 bestückt, letzterer mit 2 antiparallelen HSCH 9161. Mit meinem aufwendigen Mess-Equipment (Abb. 15), zwei WR 10 Dämpfungs-Geräten mit jeweils 50 dB und 0,1 dB Ablesung habe ich Signal/Rausch-Verhältnis-Messungen durchgeführt. Als Vergleichs-Objekt nahm ich einen von mir aufgebauten Oberwellenmischer mit der Single-Diode MA4E1317. Dabei konnte ich die guten Messergebnisse, veröffentlicht von Jürgen, DC0DA, im Dubus 2/2010 [5] mit meiner Messmethode nicht bestätigen.

Schluss

Ich weiß, dass die mechanischen Arbeiten nicht so gerne gemacht werden. In diesem Falle ist Genauigkeit aber notwendig.

Es grüßt Philipp, DL2AM, der gerne auch mechanische Arbeiten macht.

Bildunterschriften:

Abb. 1: Teile von zwei 76 GHz-Transvertern mit Wendeverstärker

Abb. 2: Fertige Transverter mit Wendeverstärker, 47 und 76 GHz

Abb. 2a: Flansch mit angedrehtem Ansatz

Abb. 3: Aufgebohrte Flansche ähnlich WG387 für WR 19, WR 12 und WR 4

Abb. 4: Konische Reibahlen und Universal-Entgrater + Gewindebohrer UNC 4/40

Abb. 5: Messen mit Schieblehre

Abb. 6: HL-Flansch, ähnlich WG387 auf Planlauf prüfen

Abb. 7: HL - WR 19 + WR 12 mit Flansch. Die Stirnseite der HL ist mit schwarzer Farbe bestrichen.

Abb. 8: Wie Foto 7 + professioneller Flansch WG387

Abb. 9: In Flansche eingepresste HL WR 19 und WR 12, fertig zum Verlöten

Abb. 10: HL - WR 12 mit Doppelflansch und aufgelegter Kupferplatte

Abb. 11: HL mit Flanschen, ähnlich WG387 und WG 383, Flanschverbindungs-Schrauben und Rundkopf-Innensechskantdreher

Abb. 12: Universal-Entgrater + Kegelfräser für HL-Adapter (Taper)

Abb. 13: HL-Reinigungs-Set

Abb. 14: Messaufbau mit 76-GHz-Filter und Isolator. Loss unter 3 dB.

Abb. 15: Messaufbau 122 GHz mit zwei 50 dB-Dämpfungsgliedern mit 1/10 dB Einteilung

Ein Vervielfacher mit der BAT15-099 Diode zur Erzeugung eines Test-Bakensignals auf 47 GHz

- auch für 24, 76, 122 und 241 GHz geeignet

von Wolfgang Demmer, DD8BD

Ursprünglich hatte ich die Absicht, einen Vervielfacher zu bauen, den ich für Test-Baken-Zwecke auf 47GHz nutzen konnte. An diesen Vervielfacher sollte u.a. eine konische Hornantenne angeschlossen werden können. Doch meine technische Ausrüstung und meine technischen Fähigkeiten reichten nicht, den für dieses Projekt entwickelten Entwurf auch in die Tat umzusetzen. Daher fragte ich Bert, PE1RKI, ob er eine Möglichkeit sähe, diesen Vervielfacher für mich zu bauen, und er nahm diese schwierige Aufgabe an!

Abb. 1: Vervielfacher x 4 von 11.772 auf 47.088 GHz

Wie man meinem Entwurf in Abb. 1 entnehmen kann, hat dieser Vervielfacher einen runden, konisch zulaufenden Hohlleiter, der sich kontinuierlich in Richtung Ausgang verjüngt. Meine Absicht war, durch die konische Struktur sowohl einen „weichen“ Übergang von der Grundfrequenz bis hin zur 47GHz-Bakenfrequenz zu bewirken, der im Vergleich zu einem sich stufenartig verjüngenden Rundhohlleiter nur geringe Fehlanpassungen aufweisen sollte, als auch eine Hochpassfilter-Wirkung zu erzeugen. Und schließlich würde man durch die Verwendung eines „geraden“ Stücks Rundhohlleiter, also ohne konische Struktur, auch ein zusätzliches Hochpassfilter an den Ausgang anschrauben können, sofern notwendig. Für eine optimale Anpassung des Eingangssignals an den Hohlleiter ist ein abstimmbarer Hohlleiterkurzschluss vorgesehen. Der Durchmesser des sich konisch verjüngenden Hohlleiters am Ausgang beträgt 4,5mm und passt somit zu dem von mir auf 47GHz verwendeten Rundhohlleiter. Nach einiger Zeit schickte mir Bert eine hervorragende technische Zeichnung, siehe Abb. 2, die zeigte, wie er gedachte, meine Idee umzusetzen. Wie aus der technischen Zeichnung ersichtlich, sollte der Hohlleiter aus vier Einzelteilen bestehen, die zusammengeschraubt werden müssten.

Abb. 2: Technische Zeichnung von PE1RKI

Schließlich entstand unter den Händen von Bert ein wunderschönes und hervorragend gearbeitetes technisches Meisterwerk!

Abb. 3: Der fertige Vervielfacher

Die Abbildung 3 zeigt den fertigen Vervielfacher mit einer gerändelten großen Feingewindeschraube als abstimmbarem Hohlleiterkurzschluss. Im Laufe der Zeit fand ich bei verschiedenen Versuchen heraus, dass dieser Vervielfacher sogar noch ein Test-Bakensignal auf 241GHz liefern kann, wenn man ihn mit 100 bis 120 mW ansteuert (solange die Diode das aushält, hi)!

Schraubt man das Endstück des Vervielfachers mit seinem Rundhohlleiter, der bei 4,5mm Durchmesser endet, ab, so hat man einen Ausgang mit 8mm Durchmesser, dem man - bei passendem LO-Signal am Vervielfachereingang - ein Oberwellensignal für 24GHz entnehmen kann. Für solche „niedrigen“ Frequenzen benötigt man wesentlich weniger LO-Pegel.

Zusammengefasst: Je nach verwendeter Frequenzaufbereitung mit ihrer bestimmten Ausgangsfrequenz und ihrem bestimmten Signalpegel (der, wenn nötig, mit einem zusätzlichen Verstärker angehoben werden kann) kann dieser Vervielfacher als Testsignalquelle für 24, 47, 76, 122 und 241 GHz verwendet werden. Dabei ist der Anschluss von konischen Hornantennen empfehlenswert.

Verfügt man über einen sehr frequenzstabilen Messender oder Synthesizer wie den HP8671B (Frequenzbereich 2 – 18 GHz) oder einen vergleichbaren Generator, mit dem ein stabiles Oszillatorsignal erzeugt werden kann, benötigt man nicht für jedes MM-Wellenband eine spezielle Frequenzaufbereitung, sondern kann das Synthesizer-Signal auf den Eingang des Multipliers geben. Zum Schutz des Synthesizers bietet es sich an, an dessen HF-Ausgang einen Isolator anzuschließen.

Auch ermöglicht die Anwendung eines solchen Synthesizers anhand eines eingestellten Frequenzversatzes eine Überprüfung der Oberwellenzahl und somit, ob man mit seinem z.B. Eigenbau-Konverter auch tatsächlich in dem beabsichtigten Amateurband gelandet ist. Die folgende Zeichnung (Abb. 4) zeigt den Anschluss der BAT-Diode an ein Stückchen Semi Rigid Koax des Typs UT141, das zuvor an eine SMA-Flanschbuchse angelötet wurde.

Abb. 4: Einsetzen der BAT15-099-Diode für den 47/76/122/241GHz-Vervielfacher

Zur Abbildung: Man lötet die Anschlüsse 3 und 4 an den Außenmantel des UT141 und die Anschlüsse 1 und 2 an den Innenleiter des Koax. Da es sich um eine antiparallele Diode handelt, geht es natürlich auch umgekehrt!

Bei meinen Versuchen hat sich klar herausgestellt, dass der Vervielfacher dann das stärkste Bakenignal liefert, wenn beide dieser in der BAT15-099 Diode befindlichen antiparallelen Dioden angelötet werden. Versuche mit nur einer der beiden antiparallelen Dioden ergaben deutlich schwächere Signale. Hier weitere Fotos des Vervielfachers:

Abb. 5: Der offene Vervielfacher

Abb. 6: Die Abbildung zeigt einen ersten mechanischen Testaufbau ohne Diode

Abb. 7: SMA-Buchse mit einseitig verlöteter Diode. Ein stärkeres Signal liefert der Vervielfacher dann, wenn beide antiparallelen Dioden angelötet sind!

Abb. 8: Beide verlöteten Dioden an der SMA-Buchse im Wellenleiter

Abb. 9: Vervielfacher mit konischen Hornantennen

In Abb. 9 sieht man in der Mitte ein kurzes rundes Stück Aluminium, das als Adapter für den Flansch einer 76GHz-Hornantenne fungiert. Mit Stücken solchen Rundmaterials lassen sich bei entsprechendem konstanten Innendurchmesser zusätzliche Hochpassfilter für verschiedene MM-Wellen-Bänder realisieren.

Ergebnisse

Beim Abgleich des Kurzschlusses mit der großen gerändelten Abstimmerschraube, die ein Gewinde mit 0,5mm Steigung hat, was ein sehr komfortables Einstellen ermöglicht, findet man Minima und Maxima. Schließt man eine andere Leitung als zuvor an den Vervielfacher an, oder ändert man die Eingangsleistung, zeigt sich, dass durch eine vorsichtige neue Einstellung dieser Abgleichschraube eine bessere Anpassung ermöglicht wird und sich der Output erhöht.

Bei Versuchen auf 47GHz mit einem DB6NT-LO (Frequenz 11,772225 GHz) zeigte sich, dass der größte Pegel aus dem Vervielfacher bei ziemlich eingedrehter Abstimmerschraube (back short) zu erzielen war. Für eine zusätzliche Anpassung kann die 3mm-Abstimmerschraube gegenüber dem Einkoppelstift verwendet werden, insoweit sie nicht als Teil einer Halterung für eine z.B. Varaktordiode im Gehäusestil eines Gun-Elements dient. An den 47-GHz-Flansch des Vervielfachers kann die bereits früher beschriebene konische Hornantenne angeschlossen werden. Natürlich lassen sich auch andere Antennen mit passendem Flansch oder mit zusätzlichem Adapter anschließen. Mittlerweile habe ich auch einen Adapter für 76GHz und ein konisches Horn für 76GHz, die alle zu dem Vervielfacher passen und problemlos angeschraubt werden können.

Um die Leistungsfähigkeit dieses Vervielfachers einschätzen zu können, wurde ein DB6NT-LO (Frequenz 11,772225 GHz) als Bakenoszillator verwendet. Hier die Ergebnisse eines Tests auf 47GHz, der über eine Distanz von einem Meter vorgenommen wurde:

1. Vervielfacher mit konischem Horn ohne Diode, nur mit Einkoppelstift: Signalpegel bis zu 33dB.
2. Vervielfacher mit konischem Horn und mit 1/2 BAT15-099 Diode: Signalpegel bis zu 47dB.
3. Vervielfacher mit konischem Horn und zwei angeschlossenen Dioden der BAT15-099: Signalpegel bis zu 62dB.
4. Nur DB6NT-LO (ohne Vervielfacher) mit breitbandiger Eigenbau-Vivaldi-Antenne: Signalpegel 40dB.

Bei einer Distanz von einem Meter besteht auf 47GHz eine Streckendämpfung von 66 dB. Verwendet man nur den LO mit einer Vilvaldi-Antenne, wobei sämtliche Oberwellen mit abgestrahlt werden, ergibt sich ein Budget von $66 + 40 \text{ dB} = 106 \text{ dB}$, was einer Strecke von 100 Metern entspricht. Verwendet man den Vervielfacher mit seiner Hochpassfunktion und der konischen Hornantenne (Gewinn ca. 18-20dB), erhöht sich das Budget auf 128 dB, was einer Strecke von ca. 1250 Metern entspricht.

Alternative Möglichkeiten zum Bau eines Vervielfachers

Besteht keine Möglichkeit, den oben beschriebenen Vervielfacher mit einer Drehbank zu bauen, so gibt es alternative Lösungsmöglichkeiten. Man kann z.B. einen rechteckigen Hohlleiter ohne Verjüngung der Innenmaße zusammen mit einer pyramidischen Hornantenne verwenden. Andere Möglichkeiten wären das stufenweise Aufbohren von Aluminium-Rundmaterial oder das v-förmige Aussägen eines rechteckigen Hohlleiters, um einen kontinuierlicheren Übergang von tiefer zu hoher Frequenz zu verwirklichen.

Vervielfacher

Hohlleiterübergang für Multipler von 11,772 auf 47,088 GHz
- alle Bohrungen stellen Rundhohlleiter

Material: Alu-Rundstab mit 30 oder 40mm Durchmesser
Zwischen Innenleiter und Außenleiter des UT-141 wird eine BAT15-099 eingelötet

Abb. 10

Abb. 10 zeigt meinen allerersten Entwurf für einen solchen Vervielfacher. Der erste Teil des gesamten Rund-Hohlleiters hat einen Durchmesser von 18mm, so dass abgesehen von einer für eine 47-GHz-Oberwelle geeigneten Frequenzaufbereitung auch Frequenzaufbereitungen für andere MM-Wellen-Bänder angeschlossen werden können. Ein Rundhohlleiter mit 18mm Durchmesser ist für Frequenzen von ca. 9769 bis 12722 MHz verwendbar und mit Verlusten sogar bis zu ca. 16181 MHz nutzbar.

In der Abbildung erkennt man die stufenweise Verjüngung des runden Hohlleiters, je kleiner die Wellenlänge, um so geringer der Innendurchmesser des runden Hohlleiters. Schließlich stellt jedes abgestufte Stück Hohlleiter auch ein Hochpassfilter dar.

Heute erscheint mir dieses ganze Gebilde etwas sehr kurz geraten zu sein, beim Bau dieser Variante des Vervielfachers würde ich mich hinsichtlich der Länge des Vervielfachers eher an den Maßen aus der technischen Zeichnung (Abb. 2) von Bert, PE1RKI, orientieren. Weil mir keine Berechnungsgrundlagen für den jeweiligen Durchmesser und die jeweilige Länge der einzelnen Stufen mit ihrem jeweils kleineren Durchmesser zur Verfügung standen, nahm ich an, dass bei diesem Modell Fehlanpassungen und dadurch erhöhte Verluste auftreten könnten. Daher entschied ich mich für einen zweiten Entwurf mit konisch zulaufendem Rundhohlleiter, wie oben beschrieben.

Dennoch stellt dieser Entwurf mit der gestuften Verjüngung des Rundhohlleiterdurchmessers eine Lösungsmöglichkeit für den Amateur dar, der über eine solide Tischbohrmaschine verfügt. Der Hohlleiterkurzschluss könnte vielleicht mit einer größeren Schraube realisiert werden, wobei ein Feingewinde von ganz großem Vorteil wäre, insoweit man ein entsprechendes Gewindeschneidwerkzeug hat. Aber auch ein verschiebbares Stück zylindrischen Metalls könnte verwendet werden. Hier nun eine weitere Lösungsmöglichkeit: Ein üblicher rechteckiger Hohlleiter wird sowohl auf der schmalen wie auch auf der breiten Seite v-förmig aufgesägt, anschließend in Richtung Innenzentrum zusammengebogen und schließlich zusammengelötet.

Abb. 11: Einfacher Vervielfacher aus v-förmig aufgesägtem Hohlleiter

Die allereinfachste Lösung, um für erste Tests auf einem MM-Wellen-Band ein Signal in die Luft stellen zu können, lässt sich wie in der obigen Abbildung 11 unter "Addition" (Zusatz) gezeigt mit einem Stück Semi Rigid Koax, einem SMA-Verbinder und einer BAT15-099 Diode realisieren. Hat man einen kleinen Parabolspiegel, eine parabolisch geformte Obstschale aus Metall, einen hohlen Rasierspiegel aus Metall oder gar nur eine parabolisch geformte Suppenkelle, kann man so ein Gebilde aus Semi Rigid als Parabolspiegel-Erreger einbauen und auf diese Weise noch einen gewissen zusätzlichen Antennengewinn erwarten.

Ich hoffe, einige sinnvolle Anregungen und den Anstoß zu weiteren eigenen Ideen gegeben zu haben und wünsche viel Erfolg und Spaß beim Basteln und Experimentieren!

Vy 73 Wolfgang Demmer, DD8BD - E-Mail: dd8bd@gmx.net

Applied Conversion of Segmented Wires from NEC2 to 432 MHz Yagi Elements

- for Semi-Insulated Mounting above the Boom
using Standard Insulators
(Part 2)

by Hartmut Klüver, DG7YBN

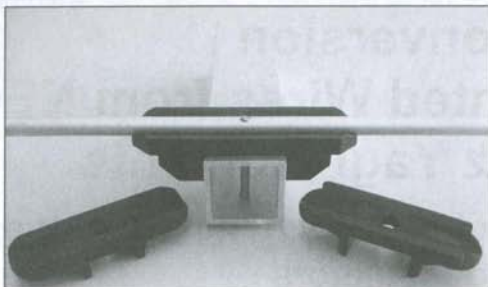
During prototyping the EF7015 antenna for production at WiMo, measurements showed it resonated several MHz too high. At first they thought that the analyser was defective! This behaviour could not be overcome by modifying DE and D1 in the usual way. The Yagi was still so much too high in frequency that the quest for a BC for the specific construction style became essential. As demonstrated for 144 MHz in [1], the NEC2 simulation can be tailored to a specific construction style. Admittedly the higher target frequency requires refining both the measuring system mechanics as well as the applied theory.

Relevance of Segmentation Density of the Design

It became evident that also on 432 MHz the segmentation density of the design must be considered for a model with an exact duplication of properties. Between an Auto-Segmentation in 'Conservative Mode' and 27 segments per wire up to 1 MHz difference may be found. Regrettably only a few designs state the segmentation density used. Only the segmentation specification that was added to the VE7BQH table when switching between EZNEC and TANT reflects the necessity for this specification. Some designers seem conscious of the fact that their models do not fit the planned resonance frequency on this band if the NEC geometry is published 1:1. Changes of several millimetres are made by implication or with different motivations. Others give geometries derived using YO or other programs. This work can not deliver on such. Those who want to appropriate the shown method when building the next Yagi, must consult its designer or type in the geometry in a NEC2 based program himself and vary the segmentation density until a match to the published data is achieved.

Specific Construction Method

Elements are secured to the boom using plastic insulators and a stainless steel M3 screw through boom and element. The elements consist of 6 mm diameter rods of AlMgSi0.5 material (EN AW6060). The M3 screw is mounted upside down, contrary to usual practice, with the head facing down and the threaded shank passing upwards through the boom. The elements have an M3 thread. This style is similar to what Andy, DJ3JJ showed in Dubus 1/2010 [2]. Three types of element insulators are used (WiMo #23040.15, #23040.20, #23040.25). They consist of Polyamide (PA 6) with an unspecified amount of carbon black to make them UV resistant. The straight, split DE is made using 10 x 1 mm tube and a 10 mm long centre insulating member in the usual way. The coax is connected in the shortest possible way (< 7 mm). Furthermore, on 432 MHz it is very important that the DE is as close to the plane of the elements as possible. An offset of several millimetres will easily lead to a shift in frequency of some 100 kHz and thus a decrease in RL of 10...15 dB. We see this very clearly with the folded dipole employing a commercial box where the FD was mounted upside down in order to make room for the box (see example EF7015). It should be understood, that the figures derived for tuning the simulation can not be transferred to other construction styles without a loss in accuracy. Especially, higher insulators and fully insulated mountings decrease the BC to a fraction; the mere change to plastic screws raises f_{res} by approximately 200 kHz. The Base BC factors given here are not transferable to other construction styles. If required, they must be scaled with the help of reference measurements.



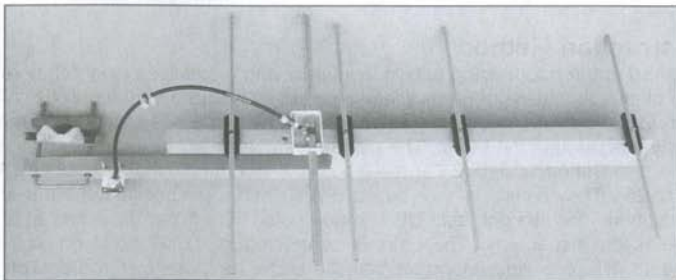
Pic. 1: Semi - insulated mounting of elements, here on 25 x 25 mm boom

Validity for Narrow Band Yagis

As in my article describing BC for 144 MHz [1] the following methods and factors apply to designs with an approximately as narrow resonance curve as the measurement Yagi described here, and the example EF7015 by Ljubisa Popa, YU7EF. Broadband designs seem to need the Base BC on that part of the complete structure that represents the higher section of the resonance curve. The part that represents the lower part of the resonance curve might, depending on design and segmentation density, result in a smaller or even negative BC. The exact figures can be determined by comparing the resonance point at a specific segmentation density against a simulation with 11 segments/wire.

Semi Professional Measurement Model

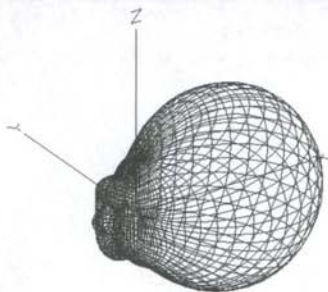
In order to increase accuracy for a good fitting procedure necessary at this higher frequency a semi professional measurement model was built. The boom consists of a profile with a cross section of 62 x 20 mm of Polypropylene (PP). All necessary fixing screws around the DE and aluminium lower boom are made of nylon. I must thank Justin, G0KSC for the suggestion of a boom completely made from plastic. But a plastic boom does not mean that there would be no influence: The dielectric constant for the PP used for the boom is 2.3 and the loss factor is 3.5. The element insulators made from PA 6 with glass fibre and furnace incorporation show a relative permittivity of 4.7 (all figures acc. IEC 60250 at 1 MHz). The influence of the stainless steel M3 screws adds to that. Then the model shall comply with the specific building style to serve as a Base Value for the fitting to the NEC simulation. The aim of the fitting process is an accurate conversion from a simulation model to this real method of construction. The elements are trimmed < 1/10 mm with a calliper gauge and straight grinding wheel. The insulator for the DE is an ABS box of 50 x 35 x 20 mm (Hammond #1551GGY). The offset from element plane to DE represents 0 mm. The balun was trimmed on the analyser and additionally fitted with ferrite cores.



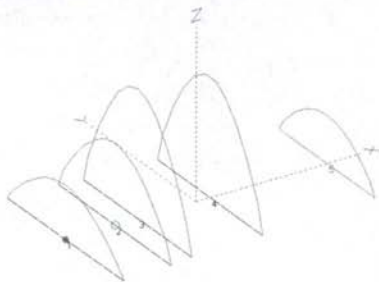
Pic. 2: Assembly of the measurement Yagi

Electrical Data of the 70-5

In an analogy of the 144 MHz tests the '70-5' design is a 5 elem. Yagi of a medium Q and relatively marked suppression of all rearward lobes. The 70-5 is designed with 11 segments per wire as well. That leads to an f_{res} of 432.20 MHz. For the further examination a stepped segmentation of 11, 11, 11, 10, 9 with an $f_{res} = 432.30$ MHz was used.



Pic. 3: EZNEC 5+ 3D - Pattern of the 70-5



Pic. 4: EZNEC 5+ Currents on the 70-5

Gain: 10.35 dBi
 F/B: 23.95 dBi
 Impedance: (49.44 - 0.66) Ohm @ 432.30 MHz & 11-9 Segments

	Pos.	Ø	NEC Geom.	Boom 15x15	Boom20x20	Boom25x25
Refl.	0	6	336.0	340.8	343.5	347.0
DE	91	10	323.0	327.8	330.5	334.0
D1	133	6	311.5	316.3	323.8	322.5
D2	266	6	302.5	307.3	310.0	313.5
D3	480	6	277.0	281.8	284.5	288.0

Table 1: Geometry of the 70-5 for resonance with various boom dimensions

Frequency Dependence of Velocity Factor

The velocity factor of a coaxial cable is determined by its dielectric material. The dielectric constant or permittivity of the plastics used - PE or PTFE - is frequency dependent. For PE the dielectric constant is 2.1 at 50 Hz and 3.0 at 1 MHz according IEC 60250. The the dielectric loss factor for PTFE is 0.5 at 50 Hz and 0.7 at 1 MHz. So the established figures for shortening factors are but guiding values for the VHF/UHF range. Thus the cable should in any case be measured with a dipmeter but an analyser (transmission mode, see appendix) is to be preferred. The cable is fitted with ferrites in turn, in order to minimise an active influence of the cable braid on the driver cell.

More about NEC2 Models and Convergence

With increasing segmentation density NEC-models converge towards a limit for the calculated impedance and average gain. The initial thought to achieve a gain in model accuracy by increasing the segmentation density is understandable. Though we must realise that we are handling a complex mathematical model that replicates the real world. The convergence slopes of both lines generally do not reach their limits simultaneously. So the plausibility and accuracy of the simulation are coming to a end here. For example of the 70-5 this difference is found to be 1 MHz. The average gain reaches its limit at 15 segments at approximately 431.6 MHz whereas the impedance comes in at 29 segments and approximately 430.6 MHz. This minimum uncertainty of 0.23% we have to attribute to the mathematical model. Using stepped segmentation can lessen the difference slightly, but the magnitude persists.

For the transition from model to real-world build we must ask on what kind of construction NEC is standardised internally? In the geometry editor of NEC we enter the parameters for segment diameter and material. But excluding the quest for boom corrections, real element fixings, insulators, screws etc. we do not have access to the model. Some of the characteristics of our highly optimized Yagi may be degraded when the geometry is not tailored to the mechanical implementation.

Convergence Gradient Chart

In order to show the dependency between convergence and segmentation density I have compiled the following graph (Chart 1). It shows the convergence gradients of impedance and average gain with and without losses of the chosen element materials against segmentation density within the area of convergence.

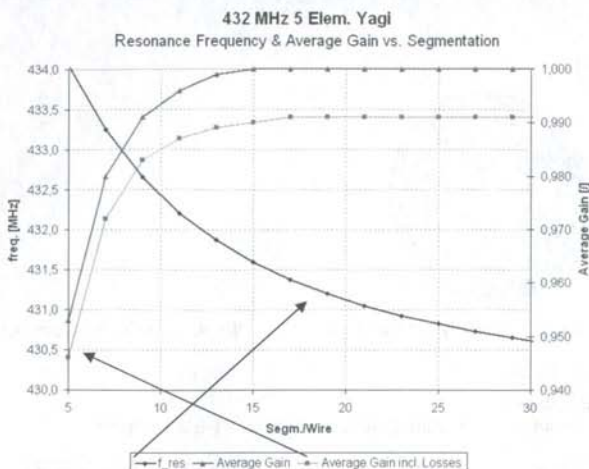
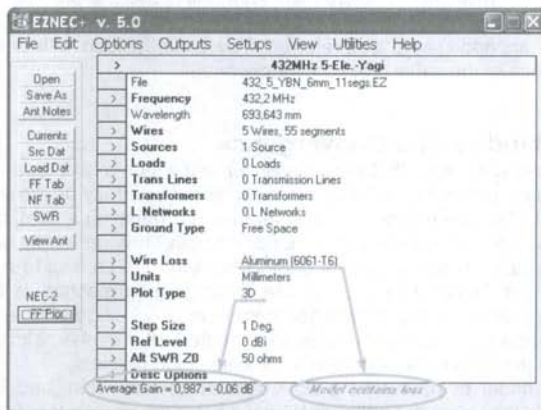


Chart 1: NEC2 simulation of the 70-5 over variation of segmentation density (material: AL 6061)

The term average gain stands for the integration of the complete 'radiation pattern power', thus the radiation power across the (directional) radiation pattern divided by the power fed in at the source. If we retrieve the whole fed in power in the pattern the ratio is 1:1 = 1.000. For an inaccuracy in the radiation pattern resulting of a lower segmentation or losses in the structure due to real materials we do find a ratio less than 1. When using real, lossy materials the convergence limits never approach 1. It reaches a limit at a lower figure since a part of the input power is consumed by losses. Thus allowing for real losses of 1...3% a limiting value of typically 0.97...0.99 results if aluminium 6061 is chosen as the real material. In EZNEC we can find the average gain on the lower edge of the control window after generating a 3D pattern and the comment 'Model contains loss' if real, lossy material has been chosen. Related convergence charts have been compiled by Leif, SM5BSZ in 1997 to show the correlation and the emerging deviations of NEC-simulations of gain and impedance versus segmentation density [3].



Pic. 5: Screenshot of EZNEC 5+ control window with 3D Pattern

The NEC2 Model of the 70-5

Whilst the average gain of the 70-5 model is already completely convergent from a segmentation density of 15 upwards, the impedance does not reach the final value of its convergence even at 35 segments per wire. But it is clear that the final value will close to about 430.3 MHz. The measurements on the 70-5 PP-model show that with a segmentation of 11 a good conformity of simulated and measured trend of resonance frequency and return loss remains.

The intended resonance frequency different by up to 200 kHz compared with the real antenna. The R.L. was verified to within 5 dB. Also the trend before and after the resonance peak where demonstrated with this R.L. offset. That shows, that no super positioning effects between balun, DE and wave guiding structure have moved the resonance into the right place and thus that these figures withstand. The difference in frequency between consistent simulation with 11 segments to the measurement and the estimated convergence lines at beyond 35 segments is $432.2 - 430.3 = 1.9$ MHz. That equals a theoretical variation of element length of approximately 2 mm.

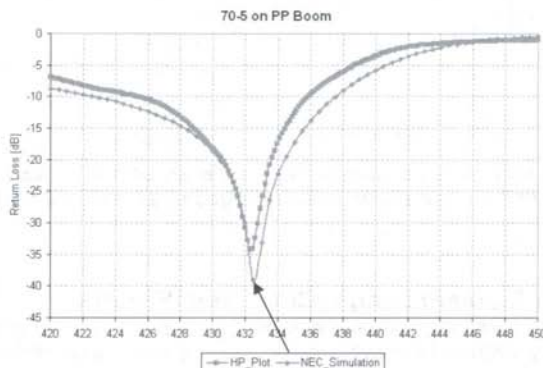


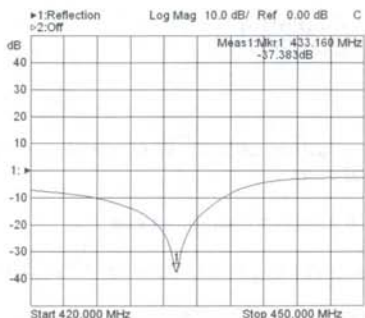
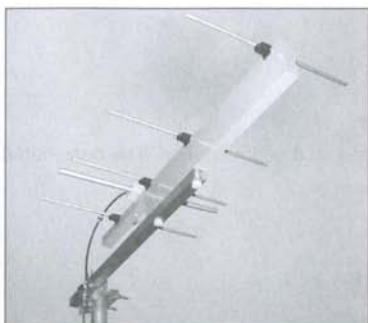
Chart 2: NEC2 simulation with 11 seg./wire and analyser plot of 70-5 with orig. geometry on PP boom

Measurements of the 70-5 Yagi with Lengthened Elements

Next to the base measurement the 70-5 on PP boom was provided with at $2.0 + 2.0 = 4$ mm and $3.5 + 3.5 = 7.0$ mm prolonged element sets and DE to begin with and measured. The outlining of BC over resonance delivered a first designation for the gradient of the straight line to plan further measurements on aluminium booms.

After these baseline measurements the PP boom was discarded; the long-winded labour of preparing further element sets and the measuring of about 100 plots on aluminium booms 15×15 mm, 20×20 mm and 25×25 mm began. For this the DE-box including the straight split DE with its balun was transferred in each case to yield comparable conditions. Consecutively three representative plot charts are displayed. The sum of measurements is represented by chart 6 'Base BC on 432 MHz'.

The measurements were carried out with a HP8710C analyser and calibrated measurement cable by Wolfgang Schmenger, DB6WY at WiMo and kindly provided for this article.



Pic. 6: 70-5 PP-Model during Measurements

Chart 3: 70-5 on 15 x 15 mm aluminium boom, BC = $2.0 + 2.0 = 4.0$ mm

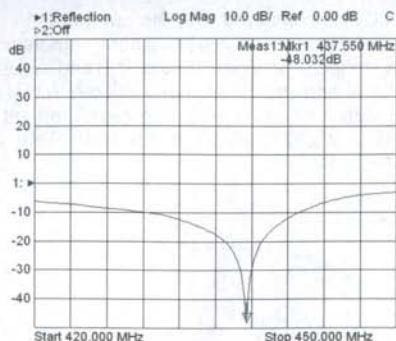


Chart 4: 70-5 on 25 x 25 mm aluminium boom, $BC = 3.5 + 3.5 = 7.0$ mm



Chart 5: 70-5 on 25 x 25 mm aluminium boom, $BC = 6.5 + 6.5 = 13.0$ mm

Engineering of the Element Length Correction Factors

All usable measurement results are summarised in chart (6). For each pairing of boom and insulator one chart line specifies the variation of length correction of the element over measured f_{res} . Zero length correction is equal to the geometry of the NEC input file.

Base BC (BBC)

The Base BC for a design segmentation density close to 11 is stated in table 2.

It encloses the difference between NEC simulation with 11 segments per wire and the element prolonging. NBC means NEC correction factor and BC stands for real BC. The NBC amounts 200 kHz here.

	PP Boom	Boom 15x15	Boom 20x20	Boom 25x25
BBC	0	4.8	7.5	11.0

Table 2: Base BC für verschiedene Boom - Dimensionen

Segmentation BC (SBC)

The averaged slope for all four lines is 1.093 MHz/mm. A deviation in the simulated resonance frequency at a higher or lower segmentation than 11 and stepped segmentation downwards is converted to a geometry change by the averaged gradient. The magnitude of the difference between the two simulation resonance frequencies depends on the convergence gradient of the particular model.

Full NEC BC (FNBC)

$FNBC = BBC + SBC$

Example

A design is carried out with 17 segments per wire and stepped segmentation. The designated boom may be 20 x 20 mm.

$BBC = 7.5$ mm; see chart 6 or table 2.

433.00 MHz (11 seg.) - 432.20 MHz (17 seg.) = 0.80 MHz

$SBC = 0.80$ MHz / (1.093 MHz/mm) = 0.73 mm

So, the 'Full NEC BC' equates:

$FNBC = 7.5$ mm + 0.73 mm = 8.23 mm

432 MHz Element Length Correction on NEC with 11 Segm. per Wire
6 mm Solid Elements, Insulator Clamps & M3 Screw with Thread in Element

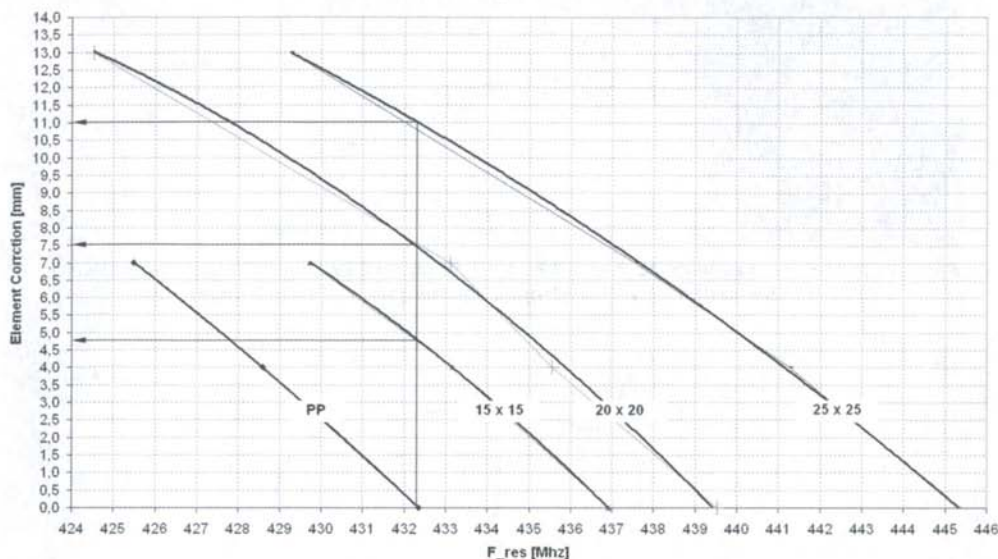


Chart 6: Base BC on 432 MHz

Discussion of the Graphs

We found quite large BC factors in relation to the original element lengths. Further the range of the minimum to maximum frequencies is 15...20 MHz. Hence we may deduce that for specific boom dimensions the line is bowed due to the fact that the whole structure would have to be scaled as well to keep up the designs full properties.

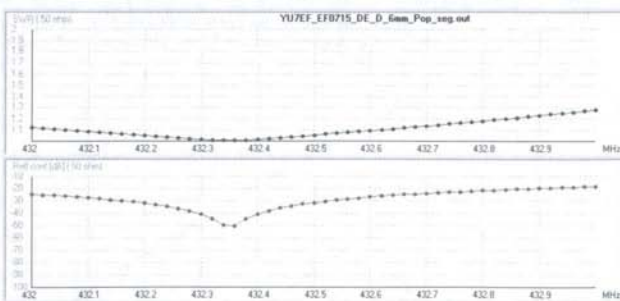
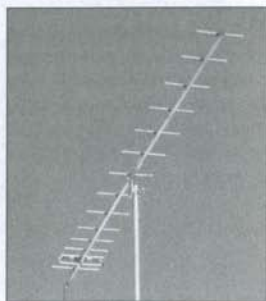
Wolfgang Schmenger, DB6WY:

"Based on the findings of the recent article on Dubus BC-factors for 144 MHz [1], we have made a series of measurements using the element holders and commonly used materials that led to the following chart (6). From the curving of the lines one clearly sees an influence of the no longer resonant structure (element distances do not fit to the F_{res} of the system). We omitted a measurement of the extrapolated element lengths on the 25mm boom (having the BC factors from the 15mm and 20mm booms). We concluded, from gut feeling and the initial results, that measurements of the PP-Yagi would not have added anything, since the element lengths and spacings would not have fitted on the PP-Yagi, and it was too cold (measurements were made in February in temperatures of 2°C). Even if these measurements had been made the result would have been a curve and not a straight line.

The beginning of curving of the lines can clearly be seen, you could go on with this until the antenna no longer works because the matching fails. In our measurements, we had usually between -20 and -40dB Return Loss.. So we reproduce only a part (the practical one) of the characteristic curves in chart 6.

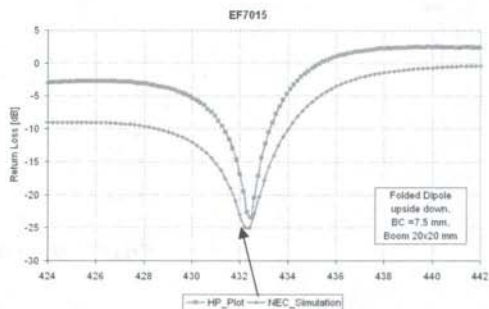
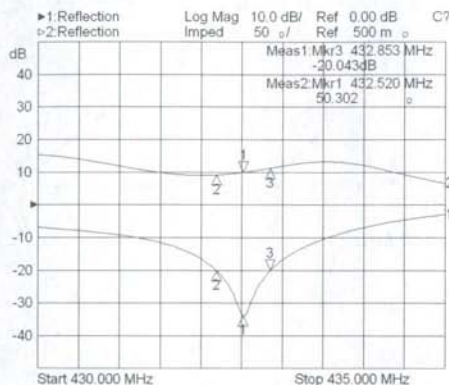
Verifying the BC using the EF7015

The EF7015 by Ljubisa Popa, YU7EF is a 15 elem.-Yagi with outstanding characteristics [4]. On a 3392 mm boom equal to an electrical length of 4.89 lambda it reaches a gain of 17.1 dBi at an antenna temperature of only 27.7 K. So the G/T is 2.71 dB. These characteristics come at a cost. Reproduction of this Yagi demands utter precision. Thus it is predestined to reassess the determined BC figures.



Pic 7: EF7015 Prototype with BC $3.75 + 3.75 = +7.5$ mm on 20 x 20 mm Boom, Photo: WiMo

Chart 7: Simulated RL with a stepped segmentation of 11 to 9 segments



Pic 8: Measured and simulated Return Loss of EF7015, construction by WiMo with folded dipole, PP holders and 7.5 mm BC.

Chart 8: Plot of EF7015 on 20 x 20 mm aluminium boom, BC = $3.75 + 3.75 = 7.5$ mm

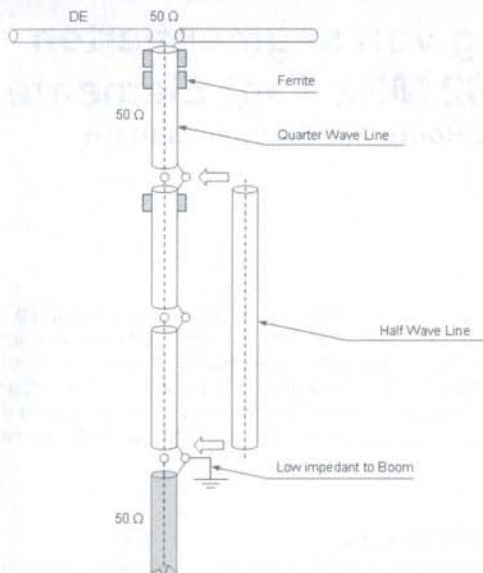
The folded dipole shown in Pic 9 is mounted upside down to be closer to the element plane. The balun was separately measured on the analyser. The best RL is reached at a little higher frequency than intended, at approx. 432.50 MHz. This is intentional and aims to make the Yagi more insensitive to rain and icing. The fine tuning of the FD of 316 mm over all length was carried out by trimming the D1. Extensive measurements on 144 MHz and 432 MHz Yagis equipped with a Folded Dipole show that against simulation results mostly D1 must be fine trimmed in length and position *). This is also true when using the nec4 engine. It is an open question whether this is the limitation of the accuracy of the simulation of an inverted folded DE with a closely coupled D1 in a 50/200 ohm system, or just the balun and DE box distorting the field. All other elements use the Base BC. It must be pointed out that the lengths given for the straight split DE on Ljubisa's website refer to same diameter as the elements for all his 432 MHz designs.

*) DB6WY: "This is clarified by the superimposed Z-line in chart 8, marked with a figure 2 on the right. This curve shows exactly 50 Ohm at marker 1 - markers 2 and 3 lower and higher Z respectively. D1 was modified for best Z and RL, but the results were several 100's kHz apart, Z = 50 Ohm at 432.100, but best R.L. was at 432.500, so this cannot be put into production."

Appendix:

Balun for Split DE at 50 Ohm

The influence of the balun in the vicinity of the driven element should be minimal. In a similar way to the tests on 144 MHz, the 70-5 measurement Yagi balun was made using a quarter wave line of thin coax. In order to cover the distance to the N-connector mounted away from the DE it was made using 3 x $\lambda/4$ lengths. The 3 consecutively arranged $\lambda/4$ pieces can alternatively be considered as a single $\lambda/4$ followed



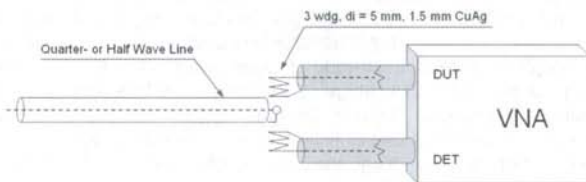
by a $\lambda/2$ line. A half wave line transfers the impedance without any change. Apart from cable losses it changes nothing.

The target resonance frequency for this to be measured as a $\lambda/4$ length thus is $432.2 \text{ MHz} / 3 = 144.067 \text{ MHz}$.

Measuring Coax Lines in Transmission Mode

A VNA's Transmission Mode is normally used for measuring filter curves or frequency responses of amplifiers. By means of the arrangement shown in Pic 10, two cables and their coupling by the specified coils an open transmission path is arranged between the DUT and DET ports. The coupling should be so loose that a disturbance is easily to be noticed. A gap of approx. 5...10 mm establishes these conditions.

Pic. 9: 3x $\lambda/4$ Balun for 50 Ohm



Pic. 10: Transmission Mode for measuring $\lambda/4$ resonances

The inserted test cable disrupts the transmission, acting as an absorption circuit. As a result at the resonance frequency of the cable as a $\lambda/4$ element, a small peak becomes visible on the swept frequency display. All cable pieces should have the frequency at which they are $\lambda/4$ calculated, since they show up as $\lambda/4$ pieces. Tend to use cables that are slightly too high in frequency. This may compensate for the connecting lengths at the soldering eyelets of the DE. If the VNA does not reach a high enough frequency, then for example a three times longer cable may be measured. The cable may be geometrically divided into three pieces afterwards.

	Quarter Wave Line	Half Wave Balun	3 x Quarter Wave Line
2 m Band	144.2 MHz	72.1 MHz	48.07 MHz
70 cm Band	432.2 MHz	216.1 MHz	144.07 MHz

Table 3: Frequency data for measuring

Literature

- [1] Klüver, Hartmut, DG7YBN: Applied Conversion of Segmented Wires from NEC2 to 144 MHz Yagi Elements for Semi - Insulated Mount above the Boom using Standard Insulators, Dubus 2/2010
- [2] Andreas Haefner, DJ3JJ: 432MHz EME with a small antenna, Dubus 1/2010
- [3] Asbrink, Leif, SM5BSZ: Compare Original NEC2 and Modified NEC2, 25.04.1997
<http://www.sm5bsz.com/antennas/neccomp.htm>
- [4] Popa, Ljubisa, YU7EF: EF7015, <http://www.yu7ef.com/ef7015.htm>

Angewandte Umsetzung von segmentierten Drähten aus NEC2 in 432 MHz Yagi Elemente für semi-isolierte Montage auf dem Boom mit Standardhaltern

von Hartmut Klüver, DG7YBN

Im Rahmen der Prototypen-Messungen für die Serienfertigung bei WiMo lag die EF7015 um einige MHz zu hoch. Zuerst glaubte man der Analyse wäre defekt! Dies Verhalten ließ sich auch auf dem üblichen Weg der Modifikation an DE und D1 nicht wirklich kaschieren. Die Yagi lag weiterhin um soviel zu hoch, dass sich die Frage nach einem nötigen BC für die relevante Bauweise geradezu aufdrängte. Wie schon für in [1] für 144 MHz absolviert, wird hier die NEC2-Simulation an die spezifische Bauweise angefitet. Allerdings ist es der höheren Zielfrequenz angemessen, sowohl die Mechanik der Messaufbauten wie auch die angewandte Theorie zu verfeinern.

Relevanz der Segmentierungsdichte des Entwurfs

Es zeigt sich auch auf 432 MHz deutlich, dass die Segmentierungsdichte des Entwurfs für einen Nachbau mit exakten Eigenschaften beachtet werden muss. Zwischen einer Auto-Segmentierung im 'Conservative Mode' und 27 Segmenten/Draht liegt bis zu 1 MHz. Leider ist die Entwurfssegmentierung bei den wenigsten Designs angegeben. Einzig die mit der Umstellung auf EZNEC und TANT hinzu gekommene Segmentierungsangabe in der VE7BQH Tabelle spiegelt die Notwendigkeit dieser Angabe.

Einige Designer scheinen sich durchaus der Tatsache bewusst zu sein, dass ihre Modelle auf diesem Band nicht ganz auf der geplanten Resonanzfrequenz liegen, wenn die NEC-Geometrie 1:1 veröffentlicht wird. So werden unter Umständen Zuschläge von einigen Millimetern stillschweigend oder mit verschiedenen Begründungen gegeben. Andere Designer geben mit YO oder anderen Programmen gerechnete Geometrien an. Dazu kann diese Arbeit keine echte Aussage treffen. Wer sich die hier gezeigte Methode zum Bau der nächsten Yagi aneignen möchte, muss den Designer befragen oder die Geometrien selber in ein NEC2-basiertes Programm eingeben und die Segmentierungsdichte variieren, bis eine Übereinstimmung zu den angegebenen Daten besteht.

Spezifische Bauweise

Diese besteht aus Befestigung der Elemente mit einem Plastikhalter und M3 Schrauben aus Edelstahl durch Element und Boom. Die Elemente bestehen diesmal aus 6 mm Vollmaterial AlMgSi0.5 (EN AW6060). Die M3-Schraube sitzt umgedreht zur üblichen Weise mit dem Kopf nach unten durch den Boom gesteckt. Die Elemente weisen M3-Innengewinde auf. Diese Bauweise ist analog zu der von Andy, DJ3JJ in Dubus 1/2010 gezeigten [2]. Es werden drei Typen von Plastikhaltern benutzt (WiMo #23040.15, #23040.20, #23040.25). Sie bestehen aus Polyamid (PA 6) mit einer nicht spezifizierten Beimischung aus Ruß, um sie UV-resistent zu machen. Der gerade, gestreckte DE ist in bekannter Bauweise mit 10 x 1 mm Rohr und 10 mm langem mittigem Isolierkörper ausgeführt. Der Anschluss des Koaxkabel erfolgt kürzest möglich (< 7 mm). Es ist auf 432 MHz von größter Wichtigkeit, dass der DE so dicht wie möglich in der Elementebene liegt. Ein Offset von mehreren Millimetern wird leicht einen Frequenzversatz von mehreren 100 kHz und Einbußen im RL von 10...15 dB zur Folge haben. Ganz deutlich sehen wir das beim Faltdipol mit kommerziellem Anschlusskästchen wo der FD zur Abhilfe kopfüber montiert wurde (s. Beispiel EF7015).

Es versteht sich, dass der gefundene Wert für das Anfiten der Simulation nicht ohne Genauigkeitsverluste auf andere Bauweisen angewandt werden kann. Insbesondere höhere Isolatoren und voll isolierte Montage verringern den BC auf einen Bruchteil; der Einsatz von Plastikschrauben allein hebt die f_{res} schon um 200 kHz. Die im Folgenden gelieferten Basis-BC-Faktoren sind nicht ohne weiteres übertragbar auf andere Bauweisen. Sie müssen im Bedarfsfall anhand von Referenzmessungen skaliert werden.

Bild 1: Semiisolierte Elementbefestigung, hier auf 25 x 25 mm Boom

Gültigkeit für schmalbandige Yagis

Die im Folgenden vorgestellte Vorgehensweise und die gefundenen Faktoren beziehen sich wie bereits in meinem Artikel zu BC auf 144 MHz [2] auch wieder auf Designs mit ähnlich schmalbandigem Resonanzverlauf wie die bezeugte Messantenne und die als Beispiel angeführte EF7015 von Ljubisa Popa, YU7EF. Breitbandigere Designs scheinen den Basis-BC auf dem Teil der Gesamtstruktur zu benötigen, der den höheren Teil der Resonanzwanne vertritt. Der Teil, der den tieferen Teil der Resonanzwanne vertritt, kann je nach Design und Segmentierungsdichte auch weniger bis hin zu negativen BC erfahren. Die genauen Werte können ermittelt werden, indem die Eckpunkte der Resonanz bei geplanter, also spezifischer Segmentierungsdichte gegen eine Simulation mit 11 Segmenten/Draht verglichen werden.

Semiprofessionelles Messmodell

Um die Genauigkeit der für ein gutes Anfitzen zu erreichenden Werte entsprechend der höheren Frequenz zu steigern wurde ein semiprofessionelles Messmodell aufgebaut. Der Boom besteht aus einem Profil mit Querschnitt 62 x 20 mm aus Polypropylen (PP). Alle notwendigen Halteschrauben im Bereich des DE und Aluminiumunterrohr bestehen aus Nylon. Für die Anregung zu einem komplett aus Kunststoff gefertigten Boom danke ich Justin, G0KSC. Ein Boom aus Kunststoff bedeutet allerdings nicht, dass es keinen Einfluss geben würde: Die Dielektrizitätszahl des für den Boom verwendeten PP liegt bei 2,3 und der dielektrischen Verlustfaktor bei 3,5. Die Elementhalter aus PA 6 mit Glasfaser- und Russbeimischung weisen eine relative Permittivität von 4,7 auf (Alle Angaben nach IEC 60250 bei 1 MHz). Dazu kommt noch der Einfluss der M3-Schrauben aus Edelstahl. Das Modell soll ja der spezifischen Bauweise entsprechen, um als Basiswert zum Anfitzen an die NEC Simulation zu dienen. Das Ziel des Fitting-Prozesses ist die genauere Umsetzung vom Simulationsmodell auf diese reale Bauweise. Die Elemente sind mit einer Schieblehre und Planschleifscheibe auf < 1/10tel mm justiert. Der Isolator für den DE ist ein Gehäuse 50 x 35 x 20 mm aus ABS (Hammond #1551GGY). Der Offset von Elementebene zu DE beträgt 0 mm. Das Symmetrierglied wurde am Analysator getrimmt und zusätzlich mit Ferritringen bestückt.

Bild 2: Aufbau der Messyagi

Elektrische Eckdaten der 70-5

Der Entwurf „70-5“ ist analog zu den 144-MHz-Tests eine 5-Element-Yagi mit mittlerem Q und einer relativ ausgeprägten Unterdrückung aller rückwärtigen Zipfel. Auch die 70-5 ist mit 11 Segmenten pro Draht design. Das ergibt eine f_{res} von 432,20 MHz. Zur weiteren Betrachtung wurde dann eine „Stepped Segmentation“ von 11, 11, 11, 10, 9 mit $f_{res} = 432,30$ MHz angewandt.

Bild 3: EZNEC 5+ 3D - Diagramm der 70-5

Bild 4: EZNEC 5+ Stromverlauf der 70-5

Gewinn: 10,35 dBi
F/B: 23,95 dBi
Impedanz: (49,44 - 0,66) Ohm @ 432,30 MHz & 11-9 Segmenten

	Pos.	Ø	NEC Geom.	Boom 15x15	Boom 20x20	Boom 25x25
Refl.	0	6	336.0	340.8	343.5	347.0
DE	91	10	323.0	327.8	330.5	334.0
D1	133	6	311.5	316.3	323.8	322.5
D2	266	6	302.5	307.3	310.0	313.5
D3	480	6	277.0	281.8	284.5	288.0

Tabelle 1: Geometrie der 70-5 für Resonanz bei verschiedenen Boom-Dimensionen

Frequenzabhängigkeit des Verkürzungsfaktors

Der Verkürzungsfaktor eines Koaxkabel wird von dessen Dielektrikum bestimmt. Die Dielektrizität oder Permittivität der verwendeten Kunststoffe PE oder PTFE ist frequenzabhängig. Nach IEC 60250 gilt für PE in etwa Dielektrizitätszahl 2,1 bei 50 Hz und 3,0 bei 1 MHz. Der dielektrische Verlustfaktor von PTFE beträgt 0,5 bei 50 Hz und 0,7 bei 1 MHz. Die bekannten Werte für die Verkürzungsfaktoren sind lediglich Anhaltswerte für den VHF/UHF-Bereich. Das Kabel soll daher unbedingt mit einem Dipmeter oder besser

Analysator (Transmission Mode, s. Anhang) ausgemessen werden. Das Kabel wurde wiederum mit Ferriten belegt, um einen aktiven Einfluss des Kabelmantels auf die Erregerzone zu minimieren.

Mehr über NEC2-Modelle und Konvergenz

NEC-Modelle konvergieren bezüglich der errechneten Impedanz und des Average Gain mit steigender Segmentierungsdichte auf einen Grenzwert zu. Der zunächst nahe liegende Gedanke, die Genauigkeit des Modells über eine Steigerung der Segmentierungsdichte zu erreichen, ist nachvollziehbar. Allerdings müssen wir uns hier vergegenwärtigen, dass es sich um ein komplexes mathematisches Modell zur Nachbildung der Realität handelt. Die Konvergenzverläufe beider Linien erreichen ihren jeweiligen Limes in der Regel nicht zugleich. Die Plausibilität und Genauigkeit der Simulation also finden deshalb spätestens hier ein Ende. Am Beispiel der 70-5 kann dieser Versatz mit 1 MHz beziffert werden. Der Average Gain erreicht seinen Limes bei 15 Segmenten entsprechend 431,6 MHz während die Impedanz bei 29 Segmenten bei etwa 430,6 MHz liegt. Diesen Mindestwert an Unschärfe von 0,23% müssen wir dem mathematischen Modell attestieren. Der Einsatz einer 'Stepped Segmentation' vermag den Versatz geringfügig zu mindern, die Größenordnung aber bleibt bestehen.

Für den Übergang von Modell zu realem Aufbau müssen wir uns fragen, auf welche Art von Aufbau ist NEC intern normiert? Wir setzen bei der Eingabe der Geometrie in NEC die Parameter für Segmentdurchmesser und Material. Aber abgesehen von der Frage nach Boom-Korrektur finden reale Elementfestigungen, Isolatoren, Schrauben etc. keinen Eingang in das Modell. Die mechanische Umsetzung macht also unter Umständen einen Teil der hoch optimierten Eigenschaften unserer Yagi wieder rückgängig, wenn die Geometrie nicht an die gewählte Umsetzung angefitzt wird.

Konvergenzverlauf-Diagramm

Um die Abhängigkeit der Konvergenz von der Segmentierungsdichte zu zeigen, habe ich das folgende Diagramm (1) erarbeitet. Es stellt den Konvergenzverlauf von Impedanz und Average Gain mit und ohne Verluste des verwendeten Elementmaterials über einer Varianz der Segmentierungsdichte innerhalb des Konvergenzradius dar.

Diagramm 1: NEC2 Simulation der 70-5 mit Variation der Segmentierungsdichte (Material: AL 6061)

Der Begriff Average Gain steht für eine Integration des vollständigen „Radiation Pattern Power“, also der Strahlungsleistung über das Richtdiagramm geteilt durch die an der Quelle (Source) eingespeiste Leistung. Finden wir die gesamte eingespeiste Leistung im Strahlungsdiagramm wieder, so ist das Verhältnis $1:1 = 1,000$. Bei aus einer niedrigen Segmentierung resultierenden Ungenauigkeit im Strahlungsdiagramm oder bei Verlusten innerhalb der Struktur bedingt durch reale Materialien finden wir ein Verhältnis kleiner 1 vor. Für die Verwendung von realen, verlustbehafteten Materialien strebt der Limes der Konvergenz nie gegen 1. Er bleibt bei einem kleineren Wert stehen, denn ein Teil der eingespeisten Leistung wird in Verlusten verbraucht. So ergibt sich entsprechend realer Verluste von 1..3% ein Grenzwert von typisch etwa 0,97...0,99, wenn Aluminium 6061 als reales Material gewählt wird. In EZNEC finden wir nach dem Erzeugen eines 3D-Pattern am untersten Rand des Kontrollfensters die Angabe zum Average Gain und die Meldung „model contains loss“ wenn reales, verlustbehaftetes Material gewählt wurde. Ähnliche Konvergenzdiagramme hat Leif, SM5BSZ bereits 1997 aufgestellt, um den Zusammenhang und die auftretenden Abweichungen in der NEC-Simulation von Gewinn und Impedanz über Segmentierungsdichte zu zeigen [3].

Bild 5: Screenshot des EZNEC 5+ Kontrollfensters bei 3D-Diagramm

Das NEC2-Modell der 70-5

Während der Average Gain bei dem Modell der 70-5 bereits ab einer Segmentierungsdichte von 15 vollständig konvergent ist, erreicht die Impedanz noch nicht einmal bei 35 Segmenten/Draht den Endwert der Konvergenz. Es ist aber erkennbar, dass der Endwert bei etwa 430,3 MHz liegen dürfte. Die Messung am PP-Modell der 70-5 zeigt, dass bei einer Segmentierung von 11 eine gute Übereinstimmung von simuliertem und gemessenem Verlauf von Resonanzfrequenz wie Return Loss besteht.

Die geplante Resonanzfrequenz wurde im realen Aufbau auf ca. 200 kHz getroffen. Das R.L. wurde dabei bis auf 5 dB nachgewiesen. Auch der Verlauf vor und nach der Resonanz wurde mit diesem Offset an R.L. gut nachgemessen. Das zeigt, dass hier keinerlei superpositionierte Effekte zwischen

Symmetrierglied, DE und Wellenleitstruktur die Resonanz an die richtige Stelle gerückt haben und diese Werte somit Bestand haben. Der Unterschied in der Frequenz zwischen übereinstimmender Simulation mit 11 Segmenten zur Messung und geschätztem Limes der Konvergenz jenseits von 35 Segmenten beträgt $432,2 - 430,3 = 1,9$ MHz. Das entspricht einer theoretischen Variation der Elementlänge von in etwa 2 mm.

Diagramm 2: NEC2 Simulation mit 11 Seg./Draht und Analyser-Plot der 70-5 mit orig. Geometrie auf PP Boom

Ausmessen der 70-5 Yagi mit verlängerten Elementen

Die 70-5 wurde nach der Basismessung zunächst auf PP-Boom mit um $2,0 + 2,0 = 4$ mm und $3,5 + 3,5 = 7,0$ mm verlängerten Elementsätzen und DE versehen und ausgemessen. Der Auftrag von BC über die Resonanzfrequenz liefert eine erste Angabe für die Steigung der Geraden, um die weiteren Messungen auf Aluminiumboom zu planen.

Bild 6: 70-5 PP-Modell während der Messungen

Nach diesen Basismessungen hat der PP-Boom ausgedient; die langwierige Arbeit der Erstellung weiterer Elementsätze und das Ausmessen von rund 100 Messplots auf Aluminiumbooms in 15×15 mm, 20×20 mm und 25×25 mm beginnt. Dazu wurden die DE-Box samt des gestreckten DE mit Symmetrierglied jeweils mit umgesetzt, um vergleichbare Bedingungen zu schaffen. Nachfolgend sind drei repräsentative Diagramme gezeigt. Die Summe der Messungen wird durch Chart 6 „Base BC on 432 MHz“ dargestellt.

Diagramm 3: 70-5 auf 15×15 mm Aluminium Boom, $BC = 2,0 + 2,0 = 4,0$ mm

Diagramm 4: 70-5 auf 25×25 mm Aluminium Boom, $BC = 3,5 + 3,5 = 7,0$ mm

Diagramm 5: 70-5 auf 25×25 mm Aluminium Boom, $BC = 6,5 + 6,5 = 13,0$ mm

Die Messungen wurden freundlicherweise von Wolfgang Schmenger, DB6WY bei WiMo mit einem HP8710C Analyser und kalibriertem Messkabel durchgeführt und mir für diesen Artikel zur Verfügung gestellt.

Entwicklung des Korrekturfaktors für die Elementlänge

Alle verwertbaren Messergebnisse sind im folgenden Diagramm (6) zusammengefasst. Für jede Paarung von Boom / Elementhalter gibt eine Diagrammlinie die Variation von Längenkorrektur des Elementes über der gemessenen f_{res} an. Null Längenkorrektur ist gleichbedeutend mit der Geometrie der NEC-Input-Datei.

Basis BC (BBC)

Der Basis-BC für eine Design-Segmentierungsdichte nahe 11 wird in Tabelle 2 angegeben. Er schließt die Differenz zwischen NEC-Simulation mit 11 Segmenten/Draht und die Elementverlängerung ein. NBC bedeutet NEC-Korrekturfaktor und BC steht für echten BC. Der NBC beträgt hier 200 kHz.

	PP Boom	Boom 15x15	Boom 20x20	Boom 25x25
BBC	0	4.8	7.5	11.0

Tabelle 2: Basis-BC für verschiedene Boom-Dimensionen

Segmentierung BC (SBC)

Die gemittelte Steilheit über alle 4 Linien beträgt 1,093 MHz/mm. Eine Abweichung in der simulierten Resonanzfrequenz bei einer höheren / niedrigeren Segmentierung als 11 und „Stepped Segmentation“ abwärts wird über die gemittelte Steilheit in eine Änderung der Geometrie umgesetzt. Die Größe des Delta zwischen den beiden simulierten Resonanzfrequenzen richtet sich nach dem Konvergenzverlauf der Resonanz des jeweiligen Modells.

Full NEC BC (FNBC)

$FNBC = BBC + SBC$

Beispiel

Ein Design ist mit 17 Segmenten/Draht und 'Stepped Segmentation' ausgeführt. Der vorgesehene Boom sei 20 x 20 mm.

BBC = 7,5 mm; siehe Diagramm 6 oder Tabelle 2.

$433,00 \text{ MHz (11 Seg.)} - 432,20 \text{ MHz (17 Seg.)} = 0,80 \text{ MHz}$

$\text{SBC} = 0,80 \text{ MHz} / (1,093 \text{ MHz/mm}) = 0,73 \text{ mm}$

So entspricht der „Full NEC BC“:

$\text{FNBC} = 7,5 \text{ mm} + 0,73 \text{ mm} = 8,23 \text{ mm}$

Diagramm 6: Basis-BC für 432 MHz

Diskussion der Graphen

Im Verhältnis zur Original-Elementlänge haben wir sehr große BC-Faktoren gefunden. Weiter beträgt die Spannweite der minimalen zu maximalen Frequenzen 15...20 MHz. Daraus können wir ableiten, dass der Linienverlauf für spezifische Boombdimensionen gebogen ist, weil die gesamte Struktur ebenfalls skaliert werden müsste, um die vollen Eigenschaften des Designs zu erhalten.

Dazu Wolfgang Schmenger, DB6WY:

"Basierend auf meinen Erkenntnissen aus dem letzten Dubus-Artikel zu BC-Faktoren für 144 MHz [1] haben wir mit den genannten Elementhaltern und den üblicherweise verwendeten Materialien eine Messreihe durchlaufen, die uns folgendes Diagramm (6) bescherte. Man erkennt in der Krümmung der Kurven deutlich einen Einfluss der nicht mehr resonanten Struktur (Elementabstände passen nicht mehr zur F_{res} des Systems). Auf eine Messung der mit hochgerechneten Elementlängen auf dem 25er Boom angewendeten Elementlänge auf dem PP-Boom haben wir verzichtet (aufgrund der BC-Kenntnisse von 15er und 20er Boom). Was wir per Bauchgefühl und der Korrekturwerte der ersten Messungen als BC-Elementlänge heraus bekommen haben, hätten wir auf der PP-Yagi noch mal messen können, aber uns war es zu kalt [gemessen wurde im Februar bei teils um 2° C...] und es bringt auch nichts, weil die Elementlängen und die Abstände der Elemente auf die Struktur der PP-Yagi nicht gepasst hätten. Aber selbst damit wäre auch eine Kurve und keine Gerade heraus gekommen. Die Ansätze der Kurven sind im Diagramm schon deutlich zu sehen, man könnte das Spiel so lange weiter treiben, bis die Antenne gar nicht mehr funktioniert, sprich die Anpassung zusammenbricht. Bei unseren Messungen hatten wir ja üblicherweise zwischen -20 und -40dB Return Loss. Aber irgendwann bricht das zusammen und es ist keine Antenne mehr. Also bilden wir mit den in Diagramm 6 gezeigten Kurven nur einen (den praktischen) Teilbereich der Kennlinie ab.

Verifizieren des BC am Beispiel der EF7015

Die EF7015 von Ljubisa Popa, YU7EF ist eine 15 Elem.-Yagi mit herausragenden Eigenschaften [4]. Auf einem Boom von 3392 mm entsprechend einer elektrischen Länge von 4,89 Lambda erreicht sie einem Gewinn von 17,1 dBi bei einer Antennentemperatur von nur 27,7 K. Damit ist ihr ein G/T von 2,71 dB besichert. Diese Werte haben einen Preis. Der Nachbau dieser Yagi fordert vollste Präzision. Damit ist sie prädestiniert zum Überprüfen der ermittelten BC Werte.

Bild 7: EF7015 Prototyp mit BC $3,75 + 3,75 = + 7,5 \text{ mm}$ auf 20 x 20 mm Boom, Photo: WiMo

Diagramm 7: Simuliertes RL bei 'Stepped Segmentation' von 11 bis 9 Segmenten

Diagramm 8: Plot der EF7015 auf 20 x 20 mm Aluminium-Boom, BC = $3,75 + 3,75 = 7,5 \text{ mm}$

Bild 8: Gemessener und simulierter Return Loss der EF7015, so wie von WiMo mit Faltdipol, PP-Elementhaltern und 7,5mm BC konstruiert.

Der in Bild 9 dargestellte Faltdipol ist überkopf montiert, um weniger weit aus der Elementebene gerückt zu sein. Der Balun wurde am Analyser von Hand vermessen. Das beste RL wird etwas höher als vorgegeben bei ca. 432,50 erreicht. Dies ist durchaus Absicht und verfolgt den Zweck, die Yagi gegen Feuchtigkeitsbelag und Eis unempfindlicher zu machen. Die Feinabstimmung auf den Faltdipol von 316 mm Spitzenweite erfolgte über feinfühliges Trimmen der Position des D1. Extensive Messungen an 144-MHz- und 432-MHz-Yagis mit Faltdipol zeigen übereinstimmend, dass der D1 entgegen der Simulation

meist in Länge wie Position im Bereich 2..5 mm angepasst werden muss*. Das gilt auch für den Einsatz des NEC4-Kern. Ob hier die Genauigkeit der Simulation bei der Errechnung des gefalteten DE in Verbindung mit dem eng gekoppelten D1 gerade bei 50/200 Ohm Systemen zu einem Ende kommt, oder schlichtweg der Balun samt DE-Box das Feld stört, blieb offen.

Alle anderen Elemente besitzen den Basis-BC. Zu beachten ist, dass die auf Ljubisas Webseite für alle 432-MHz-Designs angegebenen Längen für den gestreckten DE sich auf gleiche Durchmesser wie die Elemente beziehen

*) DB6WY: „Verdeutlicht wird das durch die in Diagramm 8 eingeblendete Z-Linie, rechts mit einer 2 markiert. Diese Kurve zeigt bei Marker 1 exakt 50 Ohm, Marker 2 und 3 niedrigeres bzw. höheres Z. Unter Betrachtung von Z und R.L. wurde D1 modifiziert, es gab dabei auch Ergebnisse, wo R.L. und Z mehrere 100 kHz auseinander lagen, Z = 50 Ohm auf 432,100, bestes R.L. aber auf 432,500, das kann man so nicht in Serie bauen.“

Anhang:

Symmetrierglied für gestreckten DE bei 50 Ohm

Der Einfluss des Symmetrierglieds auf die Erregerzone der Yagi soll minimal sein. Es wurde wie schon für die Untersuchungen auf 144 MHz für die 70-5 Messyagi mit einem Viertelwellenstück aus dünnem Koaxkabel realisiert. Um die nötige Distanz bis zu der außerhalb der Erregerzone angebrachten N-Buchse zu überbrücken, wird es hier als $3 \times \lambda/4$ -Glied ausgeführt. Die 3 hintereinander angeordneten $\lambda/4$ Stücke können alternativ als ein $\lambda/4$ -Stück mit nachfolgender $\lambda/2$ -Leitung angesehen werden. Eine Halbwellenleitung gibt die Impedanz völlig unverändert wieder aus. Abgesehen von Kabelverlusten ändert es nichts. Die angestrebte Resonanzfrequenz für das als $\lambda/4$ zu messende Stück ist demnach $432,2 \text{ MHz} / 3 = 144,067 \text{ MHz}$.

Bild 9: $3 \times \lambda/4$ Symmetrierglied für 50 Ohm

Koaxialkabel ausmessen in Transmission-Mode

Der Transmission-Mode eines VNA dient eigentlich zum Ausmessen von Durchlasskurven von Filtern oder Frequenzgängen von Verstärkern. Mittels der in Bild 10 gezeigten zwei Kabel und deren Kopplung über die angegeben Spulen wird ein offener Übertragungsweg zwischen den Toren für DUT und DET hergestellt. Die Kopplung soll so lose sein, dass eine Störung leicht bemerkbar wird. Ein Zwischenraum von etwa 5..10 mm Luft schafft diese Verhältnisse.

Bild 10: Transmission Mode zur Messung von $\lambda/4$ -Lambda-Resonanzen

Das eingefügte Testkabel stört nun als Saugkreis diese Übertragung. Als Folge wird hier bei der Resonanzfrequenz des Kabels als ein $\lambda/4$ -Lambda-Kreis ein kleiner Peak im eingestellten Frequenz-Sweep sichtbar. Alle auszumessende Kabellängen müssen dazu so in der Frequenz umgerechnet werden, dass sie als $\lambda/4$ -Lambda-Stücke erscheinen. Tendenziell sollten eher leicht zu hoch gemessene Kabel eingesetzt werden. Dies kann die Anschlusslängen der Lötösen am DE kompensieren. Wenn der VNA nicht so hoch wie gefordert messen kann, dann kann beispielsweise die dreifache Länge mit entsprechend zu erwartender Frequenz gemessen werden. Das Kabel kann dann anschließend geometrisch gedrittelt werden.

	Viertelwellen-Leitung	Halbwellen-Balun	3 x Viertelwellen-Leitung
2-m-Band	144,2 MHz	72,1 MHz	48,07 MHz
70-cm-Band	432,2 MHz	216,1 MHz	144,07 MHz

Tabelle 3: Frequenzangaben zum Ausmessen

Quellen

- [1] Klüber, Hartmut, DG7YBN: Angewandte Umsetzung von segmentierten Drähten aus NEC2 in 144 MHz Yagi Elemente für semi - isolierte Montage auf dem Boom mit Standardhaltern, Dubus 2/2010
- [2] Andreas Haefner, DJ3JJ: 432MHz EME with a small antenna, Dubus 1/2010
- [3] Åsbrink, Leif, SM5BSZ: Compare Original NEC2 and Modified NEC2, 25.04.1997
<http://www.sm5bsz.com/antennas/neccomp.htm>
- [4] Popa, Ljubisa, YU7EF: EF7015, <http://www.yu7ef.com/ef7015.htm>

The Performance of the Boxkite Yagi in Space Communications Applications

by Brian V Cake, KF2YN - Copyright © 2010. All rights reserved.

Introduction

The boxkite yagi evolved from an attempt by the author to improve the gain available from a conventional long yagi for a given boom length. The design transforms a conventional yagi from what is essentially a two dimensional to a three dimensional structure, and in so doing improves the gain available by quite a substantial amount. A side benefit is that it is possible to design an antenna that gives good performance on two bands that need not be harmonically related. The development and performance of boxkite yagis is given in detail in [1]. The nature of a long yagi restricts its use to 144 MHz and above, simply because of the word "long". Long yagis are typically several wavelengths long, and this precludes their use in most situations for 50 MHz and below.

While I was developing the boxkite, I was concerned with achieving high gain and good pattern for a given boom length. The noise performance for space communications applications was not at all a concern. However, after studying the noise performance of conventional yagis for this application, I decided to check the boxkite performance [2]. The results are I think, at the least interesting, and in some cases very good compared with contemporary low noise long yagis. In all the studies herein the noise performance is evaluated using the Tant program [3], and the standard sky and Earth noise temperatures of 200K and 1000K (144MHz) and 15K and 290K (432MHz) are used, along with an elevation angle of 30°.

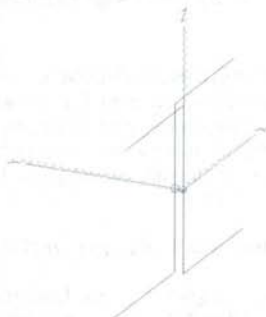


Fig. 1: Basic boxkite driven element

Boxkite yagi basics

I first described boxkite yagis back in 2004 in QEX [4, 5]. At that stage of development their performance was intriguing but not polished. Since then I have spent much time in trying to understand how to maximize performance and much of this work is described in [1]. The reader is referred to [1] for a comprehensive discussion of the technology, construction methods and performance.

The arrangement of the driven element of the antenna is shown in Figure 1. The element consists of two identical sub-elements coupled together, with the center of one of the sub-elements being the drive point. The outer horizontal wires are each approximately $\lambda/2$ long at the upper operating frequency f_3 , and the vertical spacing between the pairs of upper and lower horizontal wires is also $\lambda/2$ at f_3 . This arrangement has three resonant frequencies, f_1 , f_2 and f_3 . At f_1 and f_2 the element behaves as two over-coupled tuned circuits. The current flow at f_1 is shown in Figure 2, and at f_2 in Figure 3. At f_1 , the currents in the near-vertical wires, known as the FZ sections, are both in phase, and the current in the outer horizontal wires are in anti-phase. The resulting radiation is vertically polarized. At f_2 , the currents in the FZ sections are in anti-phase, and the currents in the outer wires are in phase. The resulting radiation is horizontally polarized. The frequency separation between f_1 and f_2 is defined by the coupling coefficient between the sub-elements, and this is controlled by the spacing between them, as well as the arrangement of the FZ sections.

The current flow at f_3 is shown in Figure 4. Here the four outer wires are driven by the FZ wires that form a pair of $\lambda/4$ open wire transmission lines. Note that there is no direct connection between the two sub-elements, but, principally magnetic coupling drives the "slave" sub-element. The two sub-elements are effectively connected in series, so the drive point impedance at the three

frequencies is quite high. The SWR sweep, in a 100Ω system, for a driven element that is designed for $f_2=144\text{MHz}$ and $f_3=432\text{MHz}$ is shown in Figure 5. Note that the ratio between f_3 and f_2 can be anywhere between about two and three, and is controlled by the coupling between the sub-elements, with tighter coupling (closer FZs) increasing f_2 and reducing f_1 while having little effect on f_3 . Methods of adjusting the coupling are shown in [1].

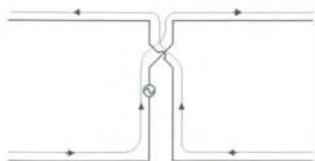


Fig. 2: Element currents at f_1

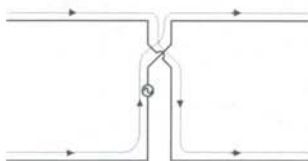


Fig. 3: Element currents at f_2

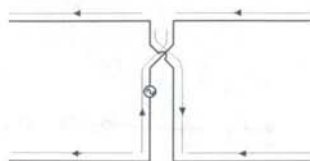


Fig. 4: Element currents at f_3

A yagi-like structure can be built using boxkite elements instead of conventional elements. The two boxkite sub-elements simply replace traditional single-wire elements, and positions and lengths are adjusted to give the desired bandwidth, gain and sidelobe levels.

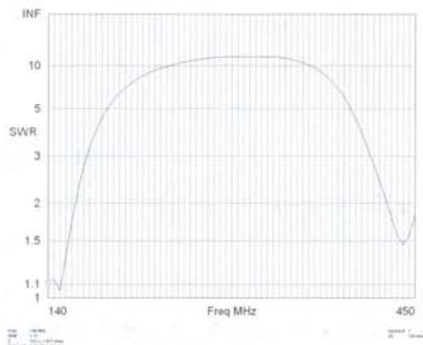


Fig. 5: SWR plot of a driven element for a dual-band 144/432MHz boxkite yagi

By choosing the correct parameters it is possible to construct boxkites that are optimized for operation at f_3 , or f_3 and f_1 (where the polarizations are crossed at f_1 and f_3), or f_3 and f_2 , where the polarizations are identical at both frequencies. For the latter case, at f_2 the elements behave like "fat" conventional elements, and can provide very wide band operation, with gain being about equivalent to that of a conventional yagi having the same boom length. At f_3 the gain is substantially higher than that of a yagi because the boxkite behaves as four stacked yagis with sub-optimal spacings in the E- and H-planes. The gain of a conventional yagi is given by:

$$G = 10 \log 9.1(L_\lambda + 0.6) \text{ dBi} \quad (1)$$

This expression is a very good representation of the gain of contemporary yagis. The gain of a boxkite yagi having a boom length of $\geq 3\lambda$ optimized for operation at f_3 is given by:

$$G = 10 \log 9.1(L_\lambda + 3.2) \text{ dBi} \quad (2)$$

and that of a dual-band boxkite at f_3 by

$$G = 10 \log 9.1(L_\lambda + 2.8) \text{ dBi} \quad (3)$$

From the above, we can see that the f_3 boxkite has a boom length advantage of 2.6λ over a yagi, and dual-band boxkite at f_3 has a 2.4λ advantage. In the two meter band, where the wavelength is 2.08m, a single-band f_3 boxkite has a boom length that is 5.4m shorter than a yagi for the same gain. For example, a 16 element K1FO yagi for two meters is 8.36m long. The gain from (1) above is 16.23 dBi, and that from simulation is 16.3 dBi. From (2) this gain can be obtained from a boxkite that is 3.05m long. However, the length would need to be slightly more than this because at 3.05m the antenna is not $\geq 3\lambda$ long. Simulation results show that a 9 element boxkite that is 4m long has a gain of 16.18 dBi. Of course, the antenna has become a three dimensional structure, rather than an (almost) two dimensional structure, and is wider and taller than a conventional yagi, but is still very manageable compared to a yagi that is over 8m long.

These same advantages in terms of boom length do of course apply to other bands, but at bands above 1296 MHz the boom length advantage becomes quite small, so unless a dual-band boxkite is required the boxkite has little to offer, except that a stacked planar array of printed boxkites behaves as a planar horn with excellent pattern, gain and SWR performance across the entire WiFi/WiMax bands from 2GHz to 7GHz. But that's another story...

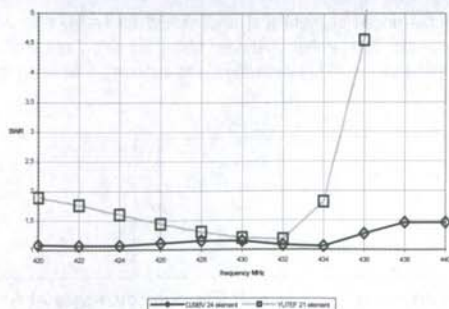


Fig. 6: SWR of DJ9BV 24 element and YU7EF 21 element yagis for the 70cm band

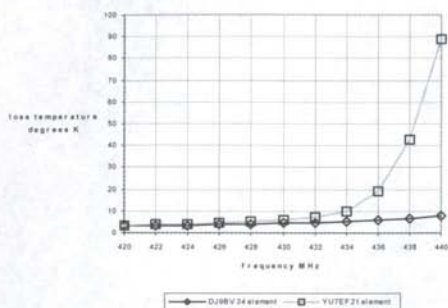


Fig. 8: Loss temperature of DJ9BV 24 element and YU7EF 21 element yagis for the 70cm band

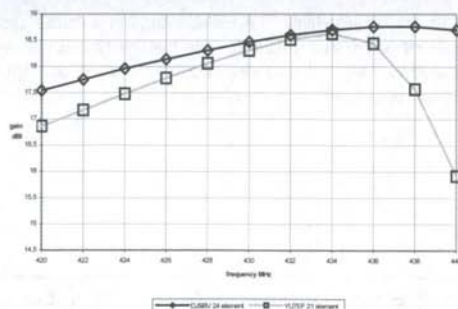
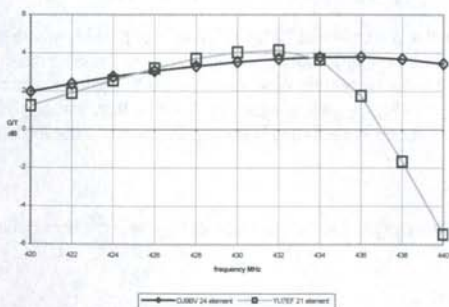


Fig. 7: Gain of DJ9BV 24 element and YU7EF 21 element yagis for the 70cm band

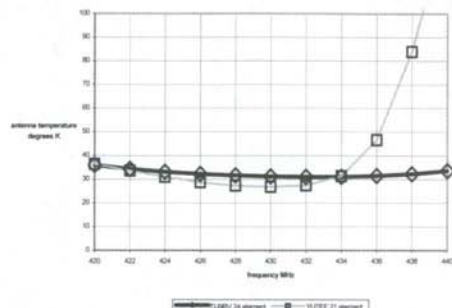


Fig. 9: Antenna temperature of DJ9BV 24 element and YU7EF 21 element yagis for the 70cm band

Comparison of the performance of boxkite yagis with that of contemporary yagis for EME

In all the following comparisons, the E-plane of the antennas is horizontal, simply to keep the amount of data within reason. However, it is important to note that physically switching or rotating the antenna polarization is becoming more popular as a way to combat the loss of signal strength associated with faraday rotation for EME work.

Fig. 10: G/T of DJ9BV 24 element and YU7EF 21 element yagis for the 70cm band

There are several conventional yagis that are specifically designed to have low sidelobes, which are important in space communications applications in order to reduce noise pickup from the Earth. In particular, the H-plane sidelobes of a conventional yagi are much worse than that of the E-plane, and can be problematic. Some yagi designs have reduced H-plane sidelobes but normally at the expense of much reduced bandwidth. The narrow bandwidth is not a problem in itself, because the antenna only needs to work over a very narrow frequency range. However, great care has to be taken in the antenna construction, in particular to make sure the simulation is accurate and, of great importance, is to be sure that the boom correction is very accurate so that the center frequency is correct.

The yagis of YU7EF and DK7ZB fall into this category. Designs by DL6WU, DJ9BV and K1FO are much broader band but are, with some exceptions, inferior in terms of antenna temperature and G/T. As an example, Figures 6 thru 10 show the performance versus frequency of a 21 element yagi by YU7EF and that of a DJ9BV 24 element yagi. Both yagis have the same boom length of 5.3m. Figure 6 shows the SWR versus frequency, and we can see that, in the 420 to 440 MHz band, there is a marked difference between the two antennas at frequencies above and below 432 MHz. Figure 7 shows that there is a similar disparity for the gain plots, with a rapid fall off in gain for the YU7EF antenna above 434 MHz. The antenna loss temperature plot for the YU7EF yagi in Figure 8 shows a rapid increase above 434 MHz. This increase has the same root cause as the increase in SWR: the feed point

impedance drops to a very low level, therefore the current required to accept a given power increases dramatically, and consequently the element $\bar{R}$ loss increases. However, as can be seen in Figures 9 and 10, the YU7EF yagi has better antenna noise temperature at 432 MHz (3.6° K quieter than the DJ9BV yagi) and a G/T that is 0.44 dB better. An accurately constructed YU7EF antenna is clearly a better choice from a G/T standpoint, but the DJ9BV design is much more forgiving of imperfect construction.

One thing we have to keep in mind in comparing antennas is that the antenna is only one part of a receiving/transmitting system. The G/T figure of merit does of course reflect the performance of the antenna in receive mode, but there are other important specifications in the receive channel path that affect the receive system performance. The two that are most important are the pre-amplifier noise figure and any losses between the antenna and the pre-amplifier. If we assume that the antenna system is designed such that the pre-amplifier is mounted at the antenna, so that its loss is unimportant, then the pre-amplifier noise performance becomes important. Ignoring this factor is akin to quoting the top speed of an automobile in vacuum... it ignores the speed limit imposed by air resistance. We will take a closer look at the effects of pre-amplifier noise later in this article.

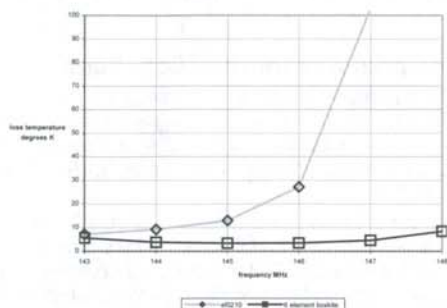


Fig. 11: SWR of a YU7EF 10 element yagi and a KF2YN 6 element boxkite for the two meter band

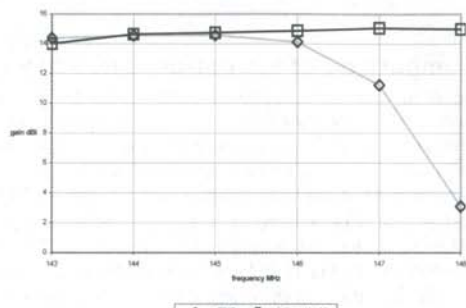


Fig. 12: Gain of a YU7EF 10 element yagi and a KF2YN 6 element boxkite for the two meter band

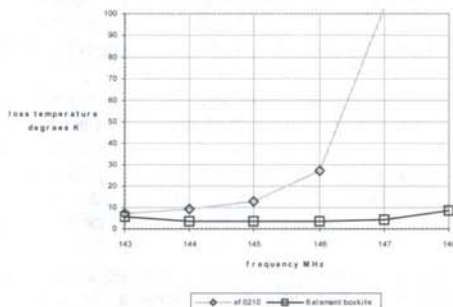


Fig. 13: Loss temperature of a YU7EF 10 element yagi and a KF2YN 6 element boxkite for the 2m band

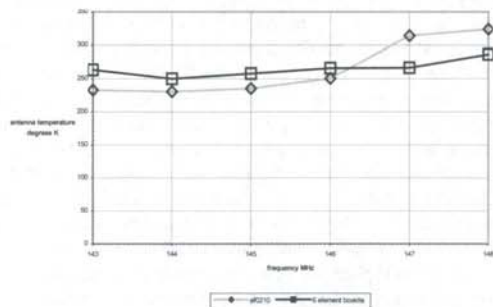
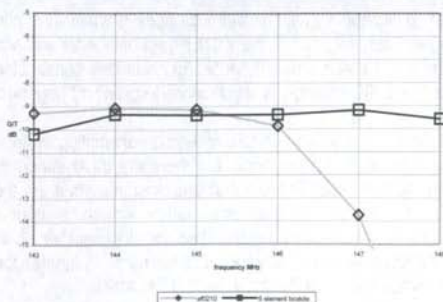


Fig. 14: Antenna temperature of a YU7EF 10 element yagi and a KF2YN 6 element boxkite for the 2m band

Comparison of the performance of yagis and boxkites for the 2m band

We will start our comparison in the two meter band. Single-yagi moonbounce stations are quite common, and have the attraction of a relatively simple single antenna, without the complications of stacking. First, we will compare the performance of a YU7EF EF0210 yagi, having a boom length of 5.4m, with that of a 6 element boxkite, with a boom length of 2.1m. They both have the same gain at 144 MHz. The plots for both antennas are in Figures 11 thru 15. The bandwidth limitations of the YU7EF antennas show up in Figures 11 and 12. In Figure 13, it is interesting to note that the antenna loss temperature of the boxkite is lower than that of the conventional yagi over most of the band. However, we have to be careful with this number. The computation of the loss temperature assumes that the simulation is very accurate, in that the Average Gain Test of the simulation gives a result of exactly 1.0 if the elements have zero loss. This is not always the case, and any error here will affect the loss temperature value. However, in the two meter band both earth and sky temperatures are so high that the loss temperature is not a major factor.

Figure 14 shows that the boxkite antenna temperature is about 20K warmer than the yagi, and Figure 15 shows a G/T that is 0.4dB in favor of the yagi. I must confess I was extremely surprised that these values were so close together, particularly since the boxkite was not designed with any of these performance criteria in mind, *plus the boxkite has a boom length that is only 40% of that of the yagi*. Part of the reason for the good noise performance is because the boxkite consists of 4 yagis, with pairs spaced



in the H-plane by about $\lambda/2$, and this is a well-known method of reducing the H-plane sidelobes that look at the Earth [2].

The baby boxkite for two meters performs very well, but what about a longer boxkite? Figures 16 thru 19 compares the data for the 9 element boxkite (boom length = 4m) with that of an EF0213 yagi (boom length = 8.1m). The above comments apply equally to these two antennas: the antenna temperature and G/T for the YU7EF antenna is marginally better than the boxkite, but again the boxkite has half the boom length of the yagi.

Fig. 15: G/T of a YU7EF 10 element yagi and a KF2YN 6 element boxkite for the two meter band

Comparison of the performance of yagis and boxkites for the 70cm band

We will continue to compare boxkite performance with that of the YU7EF yagis simply because, as far as I am aware, the YU7EF yagis represent the current state-of-the-art for noise performance. There are several ways of comparing the performance of boxkites and yagis. One is to compare antennas having the same gain, as we did in the prior section. Another is to compare the performance of antennas having equal boom length. Finally, since the "figure of merit" of antenna performance is G/T, we can compare antennas based on equal G/T, and, for simplicity, that's what we will do here.

First we look at the performance of a YU7EF EF7012 yagi that is 2.4m long. A boxkite having comparable G/T has 9 elements and is 1.3m long. This boxkite is designated as YN7009. We need a very accurate model if we are to fairly compare the performance based on noise temperature, in particular, as noted above, the Average Gain Test should return a value very close to 1.000. Boxkites need dense segmentation in order to provide an accurate model, so for all the following antennas, both yagis and boxkites, the segmentation was done using auto segmentation in EzNec at 1500MHz, and this provided exactly 1.000 for the AGT for both boxkites and yagis with zero wire loss. Note that, even though there are differences in the alignment of the peak G/T frequency in the antenna passband, I simply took the G/T at 432MHz as a comparison point when selecting comparable antennas. The data are shown in Figures 21 thru 25. From these we can see that the loss temperature for the boxkite is slightly lower than the yagi, and the antenna temperature favors the yagi by a few degrees. The boxkite gain is 0.4dB higher than the yagi, and at 432MHz the G/T of both are essentially equal, even though the boxkite has a boom length of only a little over half that of the yagi. Compared to the boxkite the SWR of the yagi climbs rapidly above 434MHz, as shown in Figure 25.

Simulations of longer boxkites were not so encouraging. The G/T of the original boxkite series and the YU7EF yagi series is plotted versus the boom length expressed in wavelengths, in Figure 26. Here we see that the length advantage of the boxkites falls off rather rapidly as boom length increases, until the EF series becomes equal or superior at 5λ and above. This was a puzzle since a plot of the gain of each series against boom length, in Figure 27, shows that the boxkite gain for any boom length is superior to that of the yagi. Also shown in this figure is a plot of the gain from (2) that I derived many years ago, that has proved accurate for boom lengths of 3λ and above, and it can be seen that the simulated boxkite data are a very close to the fit curve, so the poor G/T versus length data was not expected.

A study of the H-plane patterns of the boxkites as boom length increases showed that all the sidelobes remained low as length increased except the first forward sidelobes, which increased. An example for the 15 element boxkite is shown in Figure 28, and we can see that the first sidelobes are only 14 dB or so below the main beam. It turns out that this was a result of not-quite-optimized director lengths. In ordinary end-fire arrays, the sidelobes are very small, but in enhanced-gain Hansen Woodyard arrays the increased gain comes at the expense of good sidelobe ratio. This effect gets worse as the antenna length increases. If the phase delay along the antenna is too high for the boom length, the gain is almost unchanged but the sidelobes increase strongly. The same applies to yagis, and for a given director placement the total phase shift is controlled mostly by the director lengths. I simply reduced the director lengths to reduce the phase angle in each, and therefore the total phase shift, and this improved the sidelobe levels substantially without affecting the gain in any significant way. Note that even though the boxkites are effectively stacked in the H-plane by 0.5λ , this does not reduce the forward sidelobes to any extent. Figure 29 shows the results of "tweaking" the director lengths on the same 15 element boxkite, and we can see that the gain has dropped by 0.3dB but the first sidelobes are now improved by 3dB.

I modified all the boxkite series (except for the 9 element which doesn't have this problem because it is relatively short) and the G/T and gain results are shown in Figure 30 and 31 respectively, along with the data for YU7EF, K1FO, DJ9BV, DK7ZB yagis. We can see that the boxkite G/T is now better than or equal to that of any yagi at any boom length up to 8λ , and from Figure 31 the gain of the boxkite is substantially higher than any yagi at any boom length. However, the major advantage of shorter boom length for similar performance is much more in evidence for short booms. The boom length advantage for the boxkite, at least for the same gain as a yagi, is approximately 2.6λ for boom lengths of 3λ and upwards, but this assumes that the yagi to which it is being compared is a conventional yagi. The YU7EF series sacrifice gain for low sidelobes, and this means that the gain advantage of the boxkite on a given boom length is greater than for a conventional yagi.

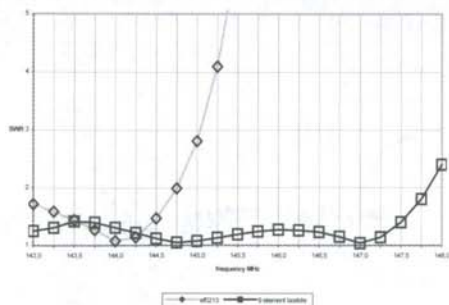


Fig. 16: SWR of a YU7EF 13 el yagi and a KF2YN 9 el boxkite for the two meter band

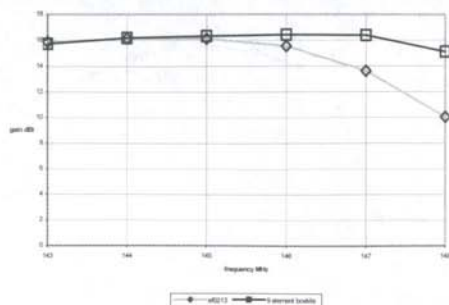


Fig. 17: Gain of a YU7EF 13 element yagi and a KF2YN 9 el boxkite for the two meter band

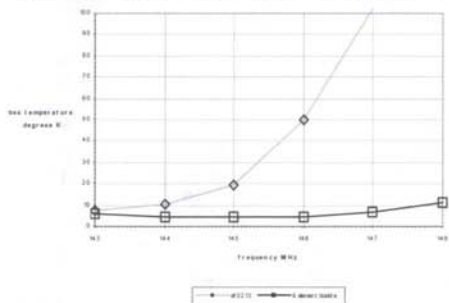


Fig. 18: Loss temperature of a YU7EF 13 el Yagi and a KF2YN 9 el boxkite for the 2m band

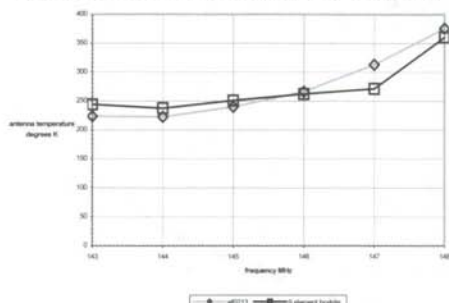


Fig. 19: Antenna temperature of a YU7EF 13 el Yagi and a KF2YN 9 el boxkite for the 2m band

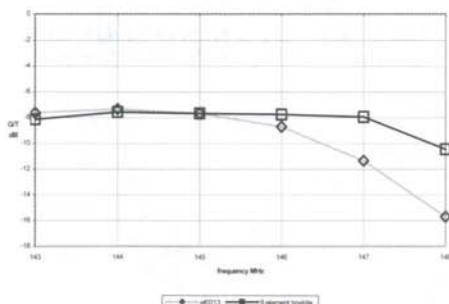


Fig. 20: G/T of a YU7EF 13 el yagi and a KF2YN 9 el boxkite for the 2m band

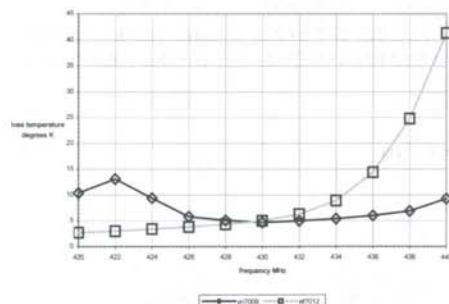


Fig. 21: Loss temperature of a YU7EF 12 el yagi and a KF2YN 9 el boxkite for the 70cm band

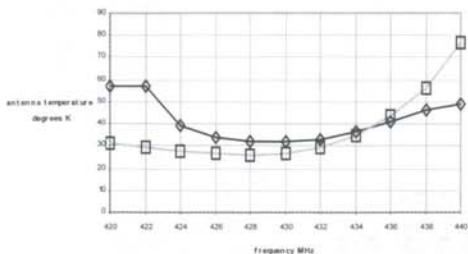


Fig. 22: Antenna temperature of a YU7EF 12 ele yagi and a KF2YN 9 el boxkite for the 70cm band
Same marking as in Fig. 21

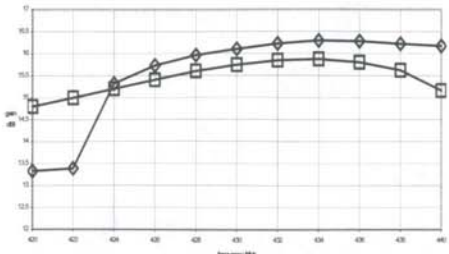


Fig. 23: Gain of a YU7EF 12 el yagi and a KF2YN 9 el boxkite for the 70cm band.
Same marking as in Fig. 21

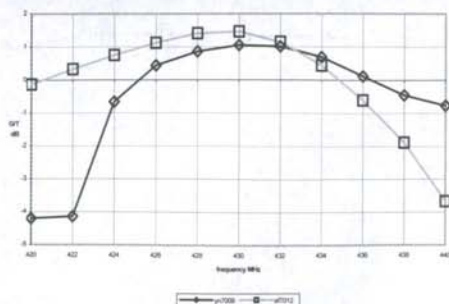


Fig. 24: G/T of a YU7EF 12 el yagi and a KF2YN 9 el boxkite for the 70cm band

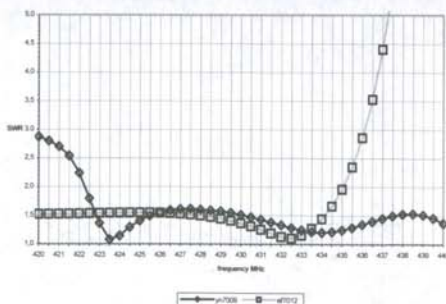


Fig. 25: SWR of a YU7EF 12 el yagi and a KF2YN 9 el boxkite for the 70cm band

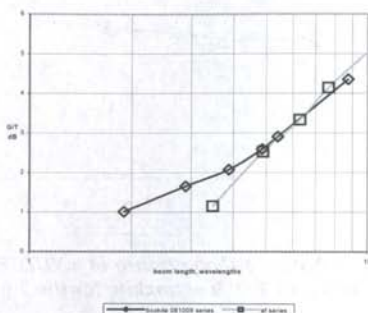


Fig. 26: G/T of the original boxkite series and the YU7EF series for the 70cm band

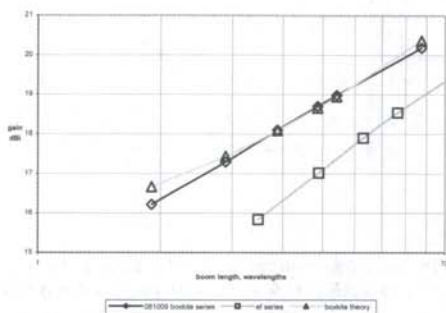


Fig. 27: Gain of the original boxkite series and the YU7EF series for the 70cm band

* Total Field

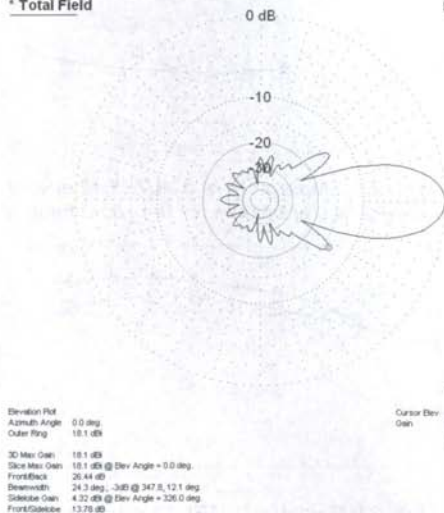


Fig. 28: Elevation pattern of a 15 element boxkite from the original series

EZNEC4 * Total Field

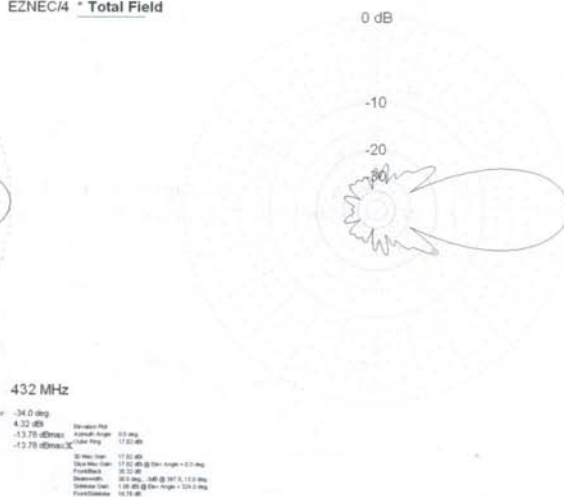


Fig. 29: Elevation pattern of a 15 element boxkite from the modified series

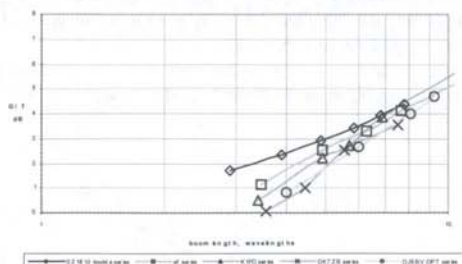


Fig. 30: G/T plots for the modified boxkite series and contemporary yagis for the 70cm band

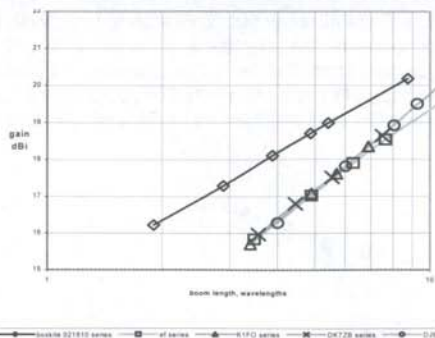


Fig. 31: Gain plots for the modified boxkite series and contemporary yagis for the 70cm band

Comparison of the performance of single-band yagi and boxkite stacked arrays

How do these antennas compare when used in stacked arrays of four for the 70cm band? We will start by comparing a stack of four YU7EF 12 element yagis, that are 2.4m long, with a stack of four 9 element boxkites that are 1.3m long. Stacking distance for the EF7012s is 1.3m in the E-plane and 1.2m in the H-plane, as recommended by YU7EF. The boxkites are spaced by 1.5m in the E-plane and 1.2m in the H-plane. This spacing is derived in the same way as regular yagi spacing from the E- and H-plane beamwidths, and seems to produce best gain and G/T.

Figures 32 and 33 show that the EF antenna is about 5° cooler than the boxkite, but the G/T for each antenna is essentially equal at 432 MHz. Figure 34 shows that the boxkite gain is 0.4 dB better than the EF. As expected, the behavior of the stacked arrays follows the same variation with frequency as that of the antennas used singly. The boxkite array is fractionally more than half the length of the yagi array.

The G/T and gain data for the 4 stack arrays of boxkites and YU7EF yagis are plotted in Figures 35 and 36, respectively. These are plots for 12, 15 and 18 element YU7EF arrays, and 9, 12, 15 and 18 element boxkite arrays. We can see that both G/T and gain for the boxkite array is higher at any boom length than the yagi. However, the real advantage of the boxkites is at short boom lengths. We will revisit the G/T performance of the two antenna series later in this article.

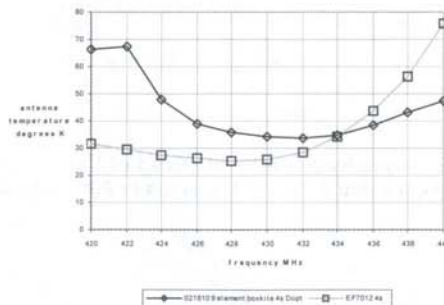


Fig. 32: Antenna temperature of stacks of 4 9 el boxkites and YU7EF 12 el yagis in the 70cm band

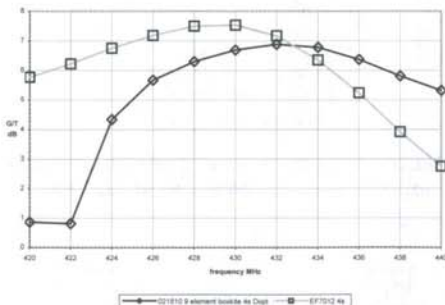


Fig. 33: G/T of stacks of 4 9 el boxkites and YU7EF 12 el yagis in the 70cm band

Comparison of the performance of single-band yagi and dual-band boxkite stacked arrays

Comparison of the performance of a dual-band boxkite array is difficult because of course there are no comparable dual-band yagi arrays, at least that I am aware of. In this case it is probably fair to compare the performance of arrays having the same boom length as the boxkite. On 2m, an array of 4 EF0206 yagis is 2.4m long and the antennas are spaced 2.6m in the E-plane and 2.3m in the H-plane. A comparably sized boxkite array is that of a 5 stack of 13 elements for 2 m and 70 cm that is slightly shorter at 2m, but has E- and H-plane spacings of the outer antennas, of 2.8m and 2.5m respectively (the reason for using a stack of 5 boxkites may be found in [1]). The data for the 2m band are shown in Figures 32 thru 35. The antenna temperature, in Figure 37 is 20K in favor of the EF, and the G/T, in Figure 38, is 0.7 dB in favor of the EF. The gain at 144 MHz, in Figure 39, is 0.4 dB higher for the EF. In the 70cm band, the comparison yagi is the EF7012 4 stack described above. Figures 40 thru 42 show

the data. Figure 40 shows that the EF stack is 11.2K cooler than the boxkite stack, but the G/T in Figure 41 is higher by about 1 dB for the boxkite. The gain of the boxkite stack, in Figure 42, is 2.4 dB higher than the yagi stack.

The foregoing shows that the dual-band boxkite array provides reasonable performance in the two meter band but is inferior to the yagi stack. Its performance in the 70cm band is much superior to the yagi. However, there are two advantages of the boxkite array in the 2m band, and these are the SWR and gain bandwidths.

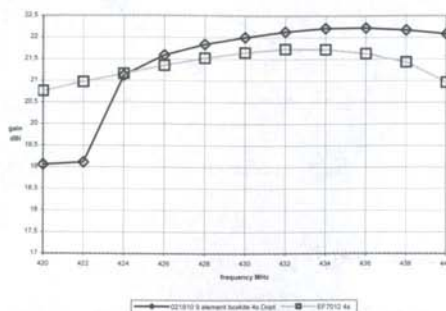


Fig. 34: Gain of stacks of 4 9 el boxkites and YU7EF 12 el yagis in the 70cm band

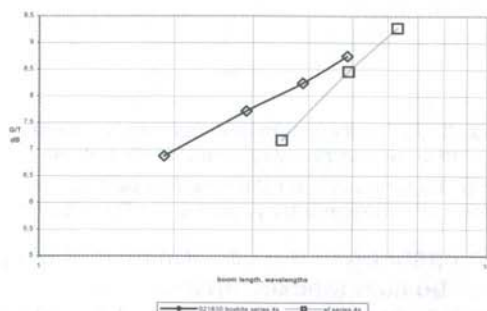


Fig. 35: G/T of stacks of 4 boxkite series and YU7EF series antennas at 432MHz

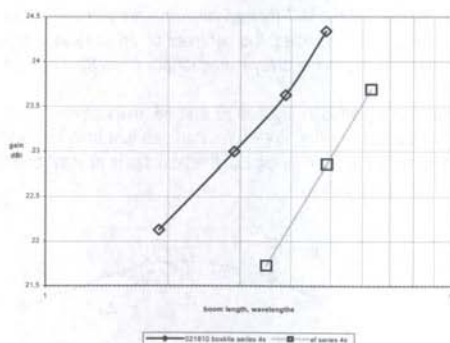


Fig. 36: Gain of stacks of 4 boxkite series and YU7EF series antennas at 432MHz

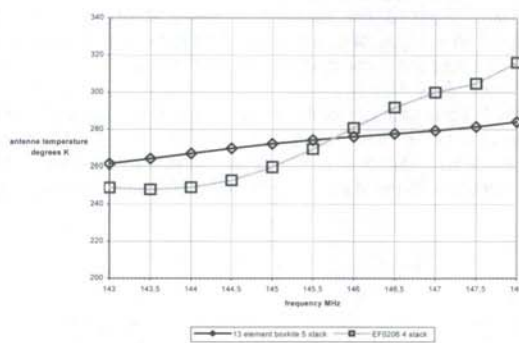


Fig. 37: Antenna temperature of stacks of 5 dual-band 13 el boxkites and 4 YU7EF 6 el yagis in the 2m band

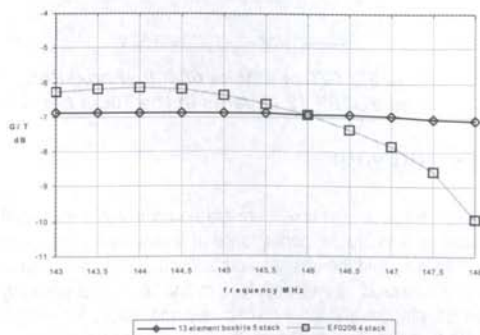


Fig. 38: G/T of stacks of 5 dual-band 13 el boxkites and 4 YU7EF 6 el yagis in the 2m band

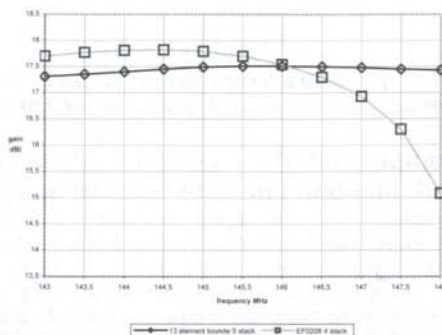


Fig. 39: Gain of stacks of 5 dual-band 13 el boxkites and 4 YU7EF 6 el yagis in the 2m band

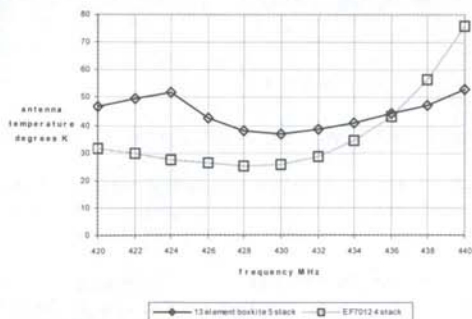


Fig. 40: Antenna temperature of stacks of 5 dual-band 13 el boxkites and 4 YU7EF 12 ele yagis in the 70cm band

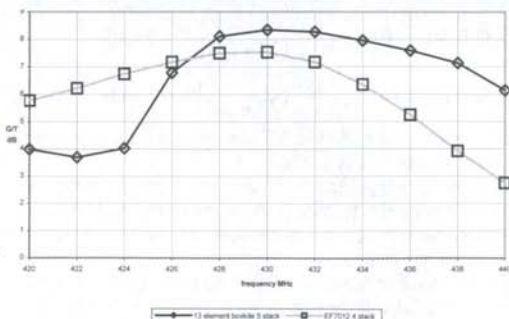


Fig. 41: G/T of stacks of 5 dual-band 13 el boxkites and 4 YU7EF 12 ele yagis in the 70cm band

The influence of pre-amplifier noise figure on practical system G/T values

As pointed out earlier, the noise performance of the preamplifier influences the achievable system G/T. In a normal low-noise antenna system that is also used for transmitting there are several noise generating mechanisms connected to the antenna. These are:

- Cable from antenna to T/R relay
- T/R relay
- Cable from T/R relay to preamplifier
- Preamplifier

If we assume that in a perfect world the cables and the T/R relay have zero loss and therefore do not generate noise, and that the gain of the preamplifier is adequate so that it and it alone determines the noise figure of the receiving system, then it is easy to calculate what the effect of the lowest available noise figure in a practical preamplifier has on the antenna system G/T.

We will assume that we are operating at 432MHz and that the preamplifier noise figure, NF , is 0.4dB, which is about the lowest noise figure currently practical for an uncooled HEMT preamplifier. The noise factor, F_n , of the preamplifier is given by

$$F_n = 10^{NF/10}$$

and the corresponding preamplifier noise temperature, T_r is

$$T_r = (F_n - 1)290 \text{ K}$$

so for our selected noise figure T_r is 28K. All we need to do to calculate the resulting G/T for our antenna and preamplifier combination is to add T_r to the antenna temperature, T_{ant} found for the antenna alone, then divide this into the numeric gain, G_n , of the antenna

$$G/T_s = 10 \log_{10} \left[\frac{G_n}{(T_r + T_{ant})} \right] \text{ dB}$$

where G/T_s is the system G/T, and $T_s = T_r + T_{ant}$

Now we will re-run the comparison of the YU7EF 21 element yagi with that of the DJ9BV 24 element design, but use G/T_s as the figure of merit, assuming a pre-amplifier noise temperature of 28K. The plots of G/T_s are shown in Figure 43, and we can see that the differences are now only 0.05dB, compared with 0.45dB for G/T for the antennas alone shown in Figure 10. This difference is now quite negligible and undetectable. As in all systems that contain multiple mechanisms that determine the overall performance, having one mechanism that outperforms the others is rarely worthwhile, especially if it might compromise another performance issue, such as pattern - and SWR - bandwidth as in this case.

Another important point is that some workers actually reduce the gain of their designs in order to improve G/T. This is not always a good thing to do from a G/T_s standpoint, as an example will show. Assume that we have an antenna, A, that has zero resistive loss and has sidelobes that are infinitely low, and that the antenna is looking at 15K sky in the 70cm band. The antenna temperature is then that of the sky.

$$T_{ant} = 15K$$

Now assume that the numeric gain, G_n , is 150 (21.76 dB), so the G/T for antenna A is $150/15 = 10$ (10dB). Next, assume we have a different antenna, B, with gain of 180 (22.5dBi) but with sidelobes and loss such that the antenna temperature is 20K. The G/T of B is then $180/20 = 9.0$ (9.5dB), and is 0.5dB inferior to that of antenna A. Now let's look at the impact of a preamplifier with a noise temperature of 28K:

From (3) above, the system G/T for antennas A and B is

$$G/T_s(A) = 10 \log_{10} \left[\frac{150}{(28+15)} \right] \text{dB} = 5.43 \text{dB}$$

$$G/T_s(B) = 10 \log_{10} \left[\frac{180}{(28+20)} \right] \text{dB} = 5.74 \text{dB}$$

So in a system the effective G/T of antenna B is better than that of antenna A by about 0.3dB, even though the "raw" G/T of A is better than that of B by 0.5dB.

The sole advantage, as far as I can see, for comparing the "raw" G/T is that it includes just the antenna itself and is not affected by any preamplifier technology improvement that might occur over time. However, it does give an incorrect comparison of actual antenna performance in a real system and can lead designers to make bad decisions when optimizing their designs, so in my view is of very little value.

Finally, we will compare the G/T_s values for the boxkite and yagi series for 70cm. These data are shown in Figure 43, and we can see that G/T_s for the boxkite series is now higher than that of any yagi for the range of boom lengths plotted up to 8λ. The plot also shows that G/T_s for the K1FO yagi series exceeds that of the YU7EF series at boom lengths of >6λ. This is probably because the latter series sacrifices gain in order to achieve high raw G/T.

Conclusions

I have presented comparative data for boxkites and yagis that is important for moonbounce and space communications applications. Even though the boxkite antennas were not designed with these applications in mind, we have seen that the performance is remarkably close to that of contemporary high-performance yagis. The dual-band boxkites in particular perform much better than I had expected, because of the compromises I had to make in the design to accommodate two bands.

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2. Brian Cake, KF2YN, "On the Reduction of Sidelobes in End-fire Arrays and Yagis", DUBUS, 3/2010
3. www.geocities.com/va3ttn/Tant.zip
4. Brian Cake, KF2YN, "Boxkite Yagis", QEX, American Radio Relay League, March 2004
5. Brian Cake, KF2YN, "Boxkite Yagis, Part 2", QEX, American Radio Relay League, September 2004

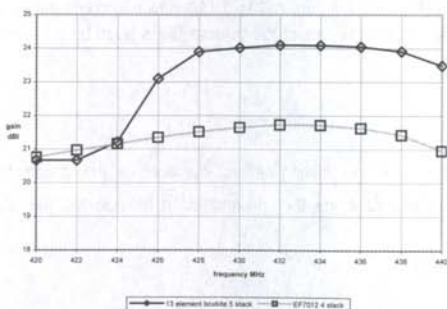


Fig. 42: Gain of stacks of 5 dual-band 13 el boxkites and 4 YU7EF 12 el yagis in the 70cm band

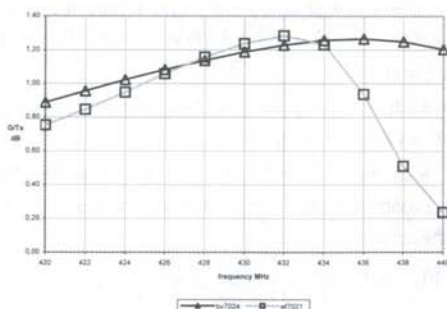


Fig. 43: G/Ts for the 24 el DJ9BV and 21 el YU7EF yagis in the 70cm band

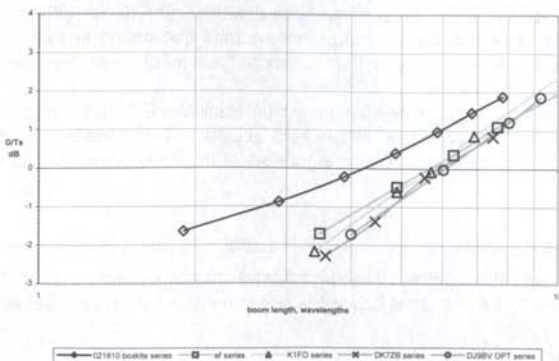


Fig. 44: G/T_s plots for the modified boxkite series and contemporary yagis for the 70cm band

Die Leistung der Boxkite-Yagi bei Weltraumapplikationen

von Brian V. Cake, KF2YN

Einführung

Beim Versuch den Gewinn einer konventionellen Langyagi bei einer gegebenen Boomlänge zu verbessern, entstand die Boxkite-Yagi. Das Design transformiert eine normale, zweidimensionale Yagi, in eine dreidimensionale Struktur. Auf diese Weise läßt sich der Gewinn um einen ordentlichen Betrag erhöhen. Ein interessanter Nebeneffekt ist die Tatsache, daß es möglich ist, die Antenne für zwei, zueinander nicht harmonische Bänder auszulegen. Der Aufbau und die erzielbaren Ergebnisse sind in [1] zu finden. Die Natur einer Langyagi beschränkt ihren Einsatz auf Frequenzen von 144 MHz und höher. Langyagis sind typischerweise mehrere Wellenlängen lang, was einen Einsatz auf 50MHz oder darunter verbietet.

Bei der Entwicklung der Boxkite-Antenne war mein vorrangiges Ziel bei gegebener Boomlänge ein sauberes Diagramm und möglichst hohen Gewinn zu erreichen. Die Rauschcharakteristik war bei diesem Einsatz für Weltraumapplikationen nicht bedacht worden. Nachdem ich aber die Rauschcharakteristik konventioneller Yagis betrachtete, entschloß ich mich, auch die Boxkite daraufhin zu untersuchen. Die Ergebnisse sind zumindest sehr interessant, und in manchen Fällen sehr gut im Vergleich zu modernen Low-Noise-Yagis. Bei allen hier gemachten Untersuchungen habe ich das Tant-Programm [3] benutzt. Die Vorgaben waren bei 144 MHz 200K für die Himmelstemperatur und 1000K für die Erdtemperatur. Bei 433 MHz nahm ich 15K für den Himmel und 293K für die Erdtemperatur an. Beide Male bezog ich mich auf eine Antennenelevation von 30°.

Grundlagen der Boxkite-Yagi

Das erste Mal habe ich die Boxkite im Jahr 2004 in der Zeitschrift QEX [4, 5] beschrieben. Dabei fand ich die Leistung faszinierend, aber die Möglichkeiten schienen mir noch nicht vollkommen ausgereizt zu sein. Seitdem habe ich viel Zeit damit verbracht, diese Ergebnisse zu verbessern. Eine umfassende Diskussion über die dahinterstehende Technik, die Konstruktionsmethoden, sowie die erzielbare Leistung einer solchen Antenne findet sich in [1]. Die Anordnung des Strahlerelements zeigt Bild 1. Das Element besteht aus zwei identischen, miteinander gekoppelten Subelementen, wobei eines davon mittig gespeist wird. Die äußeren horizontalen Drähte sind etwa $\lambda/2$ bezogen auf die obere Arbeitsfrequenz f_3 . Der vertikale Abstand der oberen und unteren horizontalen Drähte ist ebenfalls $\lambda/2$ bei f_3 . Diese Anordnung hat drei Resonanzfrequenzen, f_1 , f_2 und f_3 . Bei f_1 und f_2 verhält sich das Ganze wie ein zu stark gekoppelter Abstimmkreis. Den Stromfluß bei f_1 und f_2 zeigen die Bilder 2 und 3. Bei der Frequenz f_1 ist der Strom in den vertikalen Drähten, auch bekannt als die FZ-Sektion, in Phase, während er in den äußeren horizontalen Drähten gegenphasig ist. Die resultierende Strahlung ist vertikal polarisiert. Der Frequenzabstand von f_1 und f_2 wird sowohl durch den Kopplungskoeffizienten und damit dem Abstand der Subelemente zueinander, als auch durch Anordnung der FZ-Sektion bestimmt. Den Stromfluß bei der Frequenz f_3 zeigt Bild 4. Hier werden die vier Außendrähte von den FZ-Elementen angesteuert, die zwei offene $\lambda/4$ -Leitungen darstellen. Man beachte, daß kein direkter Kontakt zwischen den zwei Subelementen besteht, sondern die beiden „Slave“-Elemente magnetisch gekoppelt sind. Die beiden Subelemente sind effektiv in Serie geschaltet. Deshalb ist die Fußpunktimpedanz bei allen drei Frequenzpunkten recht hoch. Der SWR-Verlauf des für 100Ω ausgelegten Strahlerelements bei $f_2=144\text{MHz}$ und bei $f_3=433\text{MHz}$ zeigt Bild 5. Das Verhältnis von f_2 und f_3 kann irgendwo zwischen zwei und drei liegen. Es wird durch den Kopplungsfaktor der Subelemente bestimmt. Eine engere Kopplung erhöht f_2 und verringert f_1 , während f_3 kaum beeinflusst wird. Methoden, den Kopplungsfaktor zu beeinflussen, finden sich in [1].

Eine Yagi-ähnliche Struktur kann anstelle von konventionellen Elementen besser mit Boxkite-Elementen gebaut werden. Die beiden Boxkite-Subelemente ersetzen die herkömmlichen Elemente. Die Position und Länge wählt man so, daß die gewünschte Bandbreite, Gewinn und Nebenzipfel gegeben sind. Wählt man die passenden Parameter, ist es möglich, Boxkite-Antennen zu konstruieren, die für f_3 , oder f_3 und f_1 (mit unterschiedlicher Polarisation), oder f_3 und f_2 mit jeweils gleicher Polarisation, optimiert sind. In letzterem Fall verhält sich die Anordnung bei f_2 wie eine Antenne mit großer Bandbreite und vergleichbarem Gewinn einer Yagi mit gleicher Boomlänge. Bei f_3 ist der Gewinn substantiell höher als der einer Yagi, da sich die Boxkite wie eine vierfach gestockte Yagi mit nicht optimalem Abstand in der E- und H-Ebene verhält. Der Gewinn einer Yagi ergibt sich zu:

$$G = 10 \log 9.1(L_2 + 0.6) \text{ dBi} \quad (1)$$

Dieser Ausdruck repräsentiert sehr gut den erzielbaren Gewinn einer Yagi. Der Gewinn einer für f_3 optimierten Boxkite-Yagi mit einer Boomlänge von $\geq 3\lambda$ ergibt sich aus:

$$G = 10 \log 9.1(L_2 + 3.2) \text{ dBi} \quad (2)$$

und der einer Dualband Boxkite bei f_3 aus:

$$G = 10 \log 9.1(L_2 + 2.8) \text{ dBi} \quad (3)$$

Aus obigen Formeln sehen wir, daß sich für eine Boxkite bei f_3 eine Längenvorteil von 2.6λ gegenüber einer Yagi ergibt. Bei

einer Dualband-Boxkite ist der Faktor $2,4\lambda$. Für das 2-Meter Band bedeutet das, daß die Einband-Boxkite im Vergleich zu einer Yagi, bezogen auf den gleichen Gewinn, um 5,4 Meter kürzer ist. Beispielsweise ist eine 16-Element K1FO-Yagi für 2 Meter 8,36 Meter lang und bringt einen Gewinn von 16,23 dBi (1). Laut (2) kann ein solcher Gewinn mit einer Boxkite mit 3,05 Metern Länge erreicht werden. Allerdings muß die Antenne etwas länger sein, da die 3,05 Meter nicht $\geq 3\lambda$ sind. Die Simulationsergebnisse zeigen, daß eine 9-Element Boxkite mit 4 Metern Länge einen Gewinn von 16,18 dBi erzielt. Natürlich ist dies dann eine 3-dimensionale Anordnung und diese ist breiter und höher als eine normale Yagi. Sie ist aber im Vergleich zu einer 8 Meter langen, normalen Yagi noch gut zu handhaben.

Die gleichen Längenvorteile gelten auch für andere Bänder. Allerdings wird dieser Vorteil bei Bändern über 1296 MHz immer geringer, so daß die Boxkite hier nur wenig bietet. Außer man sucht eine Antenne mit exzellentem Strahlungsprofil, gutem Gewinn und SWR über den Frequenzbereich von 2 bis 7 GHz. Aber das ist eine andere Geschichte...

Vergleich von Boxkite-Yagis mit konventionellen Yagis für EME-Betrieb

Um die entstehende Datenmenge überschaubar zu halten, stellt bei allen nun folgenden Vergleichen die E-Ebene die Horizontale dar. Wenngleich das Ändern der Polarisationssebene durch Umschalten oder Drehen immer populärer wird, um den Signalverlusten durch die Faraday-Drehung entgegenzuwirken. Es gibt einige konventionelle Yagis, die auf geringe Nebenzipfel hin ausgelegt sind. Ein Umstand, der bei Weltraumkommunikations-Applikationen wichtig ist, um das durch die Erde verursachte Rauschen gering zu halten. Besonders die Nebenzipfel in der H-Ebene einer herkömmlichen Yagi können problematisch sein. Einige Yagi-Designs haben in der H-Ebene reduzierte Nebenzipfel. Allerdings geht das dann auf Kosten einer stark verringerten Bandbreite. Die schmale Bandbreite stellt für sich genommen noch kein Problem dar, denn die Antenne muß ja nur in einem schmalen Frequenzband funktionieren. Allerdings muß man bei der Konstruktion große Vorsicht walten lassen und die Boomkorrektur sehr genau ausführen, um dabei die Mittenfrequenz nicht zu verändern.

Die beiden Yagis von YU7EF und DK7ZB fallen in diese Kategorie. Entwicklungen von DL6WU, DJ9BV und K1FO haben eine weit größere Bandbreite und sind, mit wenigen Ausnahmen, schlechter hinsichtlich Antennentemperatur und G/T. Beispielsweise zeigen die Bilder 6 bis 10 verschiedene Parameter der 21-Element YU7EF und der 24-Element von DJ9BV jeweils auf die Frequenz bezogen. Beide Antennen haben eine Boomlänge von 5,3 Meter. Bild 6 zeigt den SWR-Verlauf im Vergleich zur Frequenz und wir sehen, daß im UHF-Bereich zwischen den Antennen ein markanter Unterschied, sowohl unterhalb als auch oberhalb von 432 MHz besteht. Bild 7 zeigt einen ähnlichen Unterschied im Gewinnverlauf mit einem rapiden Gewinnabfall oberhalb von 434 MHz bei der YU7EF. Die Antennenverlusttemperatur der YU7EF steigt simultan dazu oberhalb von 434 MHz steil an. Dieser Anstieg und der Anstieg des SWR haben die gleiche Ursache: die Speisepunktimpedanz sinkt auf einen sehr niedrigen Wert. Deshalb steigt der Strom bei gegebener Leistung dramatisch an und konsequenterweise erhöht sich auch der Elementverlust. Aus den Bildern 9 und 10 kann man sehen, daß die YU7EF-Yagi eine um 3,6K geringere Antennenrauschktemperatur und ein um 0,44 dB besseres G/T Verhältnis im Vergleich zur DJ9BV-Yagi hat. Eine möglichst exakt konstruierte YU7EF-Antenne ist klar die bessere Wahl hinsichtlich des G/T. Allerdings verzeiht eine nicht optimal konstruierte DJ9BV etwaige Konstruktionsfehler wesentlich besser.

Eines müssen wir bei den Vergleichen verschiedener Antennen aber immer bedenken: Die Antenne ist nur ein Teil eines Send-/Empfangssystems. Das G/T-Verhältnis spiegelt zwar schon die Empfangsleistung einer Antenne wider, aber es gibt daneben noch andere wichtige Parameter, die die Gesamtsystemleistung beeinflussen. Die beiden wichtigsten sind der Rauschfaktor des Vorverstärkers und die Verluste zwischen Antenne und Vorverstärker. Wenn wir annehmen, daß der Vorverstärker direkt an der Antenne montiert ist, bleibt nur noch dessen Rauschfaktor als wichtige Größe übrig. Ihn zu ignorieren, wäre so, als würde man die Maximalgeschwindigkeit eines Autos im Vakuum angeben. Die Einflüsse des Vorverstärker-Rauschens werden wir später noch näher betrachten.

Vergleich von Yagi und Boxkite für das 2-Meter-Band

Wir wollen unseren Vergleich im 2-Meter-Band beginnen. Einzelne EME-Antennen sind recht verbreitet und haben den Vorteil der Einfachheit. Sie vermeiden die Probleme von gestockten Anordnungen. Zunächst wollen wir eine YU7EF EF0210 mit 5,4 Metern Länge mit einer 6-Element Boxkite von 2,1 Metern Länge vergleichen. Aus den Bildern 11 und 15 geht hervor, daß beide Antennen den gleichen Gewinn haben. Die Bandbreiteeinschränkungen der YU7EF-Antennen werden in den Bildern 11 und 12 deutlich. In Bild 13 zeigt sich ein interessantes Detail, nämlich, daß die Antennenverlusttemperatur einer Boxkite gegenüber einer konventionellen Yagi über weite Bereiche geringer ausfällt. Allerdings ist Vorsicht geboten. Bei der Kalkulation der Verlusttemperatur gehen wir davon aus, daß die Simulation sehr genau erfolgt und der durchschnittliche Gain-Test der Simulation bei angenommenen verlustlosen Elementen exakt 1,0 ergibt. Das ist aber nicht immer der Fall und jede Ungenauigkeit hier beeinflusst den Wert der Verlusttemperatur. Im 2-Meter-Band sind allerdings Erd- und Himmelstemperatur so hoch, daß die Verlusttemperatur in diesem Fall keine so große Rolle spielt.

In Bild 14 sehen wir, daß die Antennentemperatur der Boxkite-Yagi um etwa 20K wärmer, als die der Yagi ist. In Bild 15 zeigt sich ein um 0,4 dB besserer G/T-Wert zugunsten der Yagi. Ich war sehr überrascht, daß die Ergebnisse beider Antennen so nah beieinander lagen. Zum Einen war die Boxkite in keiner Weise in Bezug auf die vorgenannten Leistungskriterien hin entwickelt worden, zum Anderen hatte die Boxkite einen um 40% kürzeren Boom im Vergleich zur Yagi. Eine Ursache für die gute Rauschcharakteristik ist, daß die Boxkite aus 4 Yagis besteht, die paarweise im Abstand von $\lambda/2$ angeordnet sind. Das ist eine bekannte Methode um die Nebenzipfel in der H-Ebene zu reduzieren [2].

Diese kleine Boxkite funktioniert sehr gut. Aber was ist mit einer längeren Version einer solchen Antenne? Die Bilder 11 bis 15 stellen die Daten einer 4 Meter langen 9-Element-Boxkite denen einer EF0213 mit 8,1 Metern Länge gegenüber. Die oben gemachten Bemerkungen gelten gleichermaßen für die beiden Antennen. Antennentemperatur und G/T einer YU7EF sind marginal besser als die einer nur halb so langen Boxkite.

Vergleich von Yagi und Boxkite für das 70cm-Band

Da meines Wissens nach die YU7EF zur Zeit das Nonplusultra in Sachen minimales Rauschen darstellt, wollen wir diese weiterhin als Referenz benutzen. Es gibt verschiedene Möglichkeiten Yagis und Boxkites miteinander zu vergleichen. Eine davon ist die, Antennen mit gleichem Gewinn gegenüberzustellen, was wir auch bisher gemacht haben. Eine andere Möglichkeit ist Antennen mit gleicher Boomlänge zu vergleichen. Da die Güte einer Antenne vor allem durch den G/T-Parameter ausgedrückt wird, wollen wir das auch hier tun.

Zunächst sehen wir uns die Leistung einer EF7012-Yagi mit 2,4 Metern Länge an. Eine Boxkite mit vergleichbarem G/T hat 9 Elemente und ist 1,2 Meter lang. Die Boxkite nennen wir YN7009. Wollen wir die Rauschtemperatur der beiden Antennen vergleichen, brauchen wir ein sehr genaues Modell, insbesondere sollte der Gain-Test ein Ergebnis nahe 1,000 ergeben. Um die Boxkites genau zu modellieren, bedarf es einer möglichst engen Segmentierung. Deshalb haben wir in EZNEC die Funktion "auto segmentation" bei 1500 MHz benutzt. Das ergibt bei angenommenen verlustlosen Elementen einen durchschnittlichen G/T von exakt 1,000, sowohl für die Boxkite als auch für die Yagi. Obwohl es Unterschiede hinsichtlich der maximalen G/T-Frequenz zwischen den Antennen gibt, nehmen wir zum Vergleich nur den G/T-Wert bei 432 MHz. Die errechneten Ergebnisse hierzu zeigen die Bilder 21 bis 25. Daraus sehen wir, daß bei der Boxkite die Verlusttemperatur und bei der Yagi die Antennentemperatur jeweils niedriger ist. Der Gewinn der Boxkite ist um 0,4 dB höher, bei gleichem G/T, und das obwohl die Boxkite nur etwas mehr als halb so lang im Vergleich zur Yagi ist. Außerdem zeigt Bild 25, daß das SWR der Yagi über 434 MHz rapide ansteigt. Simulationen längerer Boxkites waren nicht sehr vielversprechend. Ein G/T-Vergleich von Original-Boxkite und YU7EF in Abhängigkeit von der Boomlänge findet sich in Bild 26.

Daraus sehen wir, daß der Vorteil der Boxkite mit zunehmender Boomlänge recht schnell abnimmt, außer die YU7EF ist 5λ oder darüber. Das war verwirrend, da ein Plot der beiden Serien über unterschiedliche Boomlängen (Bild 27) zeigt, daß der Gewinn der Boxkite im Vergleich zur Yagi bei jeder Boomlänge höher ausfällt. Der ebenfalls im Bild dargestellte Gewinn-Plot aus (2) hat sich bei Boomlängen von 3λ und mehr als sehr genau erwiesen. Die schlechten G/T-Werte im Vergleich zur Boomlänge waren so nicht erwartet worden. Eine Untersuchung der H-Ebene zeigt, daß die Nebenzipfel über alle Boomlängen hinweg, abgesehen von den ersten Nebenzipfeln, welche größer werden, klein bleiben. Ein Beispiel einer 15-Element Boxkite ist in Bild 28 zu sehen. Dort sehen wir, daß die ersten Nebenzipfel nur etwa 14 dB kleiner im Vergleich zur Hauptstrahlrichtung sind. Es stellte sich heraus, daß der Grund eine nicht optimierte Länge der Direktorelemente war. Bei normalen endgespeisten Yagis sind die Nebenzipfel sehr klein. Bei gewinnoptimierten Hansen-Woodyard-Arrays dagegen erreicht man den höheren Gewinn auf Kosten sich verschlechternder Nebenzipfel Eigenschaften. Je länger die Antenne umso negativer macht sich dieser Effekt bemerkbar. Wenn die Phasenverzögerung im Vergleich zur Boomlänge zu hoch ist, bleibt der resultierende Gewinn nahezu gleich, die Nebenzipfel hingegen wachsen rapide an. Gleiches gilt auch für Yagi-Antennen. Bei gegebenem Direktor wird die Phasenverschiebung hauptsächlich durch die Direktorlänge bestimmt. So reduzierte ich einfach die Direktorlängen, sprich den Phasenwinkel und damit auch die Phasenverschiebung. Das verbesserte die Nebenzipfel dramatisch, ohne den Gewinn sonderlich zu beeinflussen. Obwohl die Boxkite praktisch mit $0,5\lambda$ in der H-Ebene gestockt ist, werden die ersten Nebenzipfel dadurch praktisch nicht reduziert. In Bild 29 sehen wir das Resultat der Direktorelement-Verkürzung. Der Gewinn verringerte sich lediglich um 0,3 dB, die Nebenzipfel dagegen verringerten sich um 3 dB.

Außer der 9-Element Boxkite modifizierte ich alle meine Antennen dahingehend. Die Ergebnisse im Vergleich zu den Yagis von YU7EF, K1FO, DJ9BV und DK7ZB finden sich in den Bildern 30 und 31. Wir sehen daraus, daß der G/T der Boxkite bis zu einer Boomlänge von 8λ gleich oder besser gegenüber allen Yagis ist. Der Gewinn der Boxkite ist, unabhängig von der Boomlänge, gegenüber allen Yagis substantiell höher (Bild 31). Der große Vorteil einer solchen Anordnung ist bei gleicher Leistung natürlich, wegen des kürzeren Booms, die bessere Handhabbarkeit. Der Längenvorteil der Boxkite ist bei gleichem Gewinn im Vergleich mit konventionellen Yagis bei Längen von 3λ oder mehr etwa $2,6\lambda$. Die YU7EF redziert die Nebenzipfel auf Kosten des Gewinns. Dadurch ist der Längenvorteil gegenüber der Boxkite in diesem Fall sogar noch größer.

Vergleich von Yagi und Boxkite in gestockter Anordnung

Wie verhalten sich die Antennen nun im 70-cm-Band bei 4-fach gestockter Anordnung? Wir vergleichen zunächst eine 4-fach gestockte, 2,4 Meter lange 12-Element YU7EF mit vier, 1,3 Meter langen, 9-Element Boxkites. Der Stockungsabstand der EF7012 ist, wie von YU7EF vorgeschlagen, 1,3 Meter in der E-Ebene und 1,2 Meter in der H-Ebene. Die Boxkites sind 1,5 Meter in der H- und 1,2 Meter in der E-Ebene gestockt. Diese Maße scheinen die optimalen Werte für Gewinn und G/T zu ergeben. Die Bilder 32 und 33 zeigen, daß bei 432 MHz die G/T-Werte beider Antennen gleich und die EF7012 circa 5K kühler ist. Der Gewinn der Boxkite ist um 0,4 dB höher als der der 7012. Wie erwartet ist das Verhalten des Arrays jeweils gleich dem der Einzelantennen. Das Boxkite-Array ist dabei etwas mehr als halb so lang wie das der Yagi. Gewinn und G/T-Wert beider Arrays zeigen die Bilder 35 und 36. Und zwar für 12, 15 und 18-Element YU7EF-Arrays sowie 9, 12, 15 und 18-Element Boxkite-Arrays. Wir sehen, daß sowohl G/T als auch Gewinn der Boxkite-Arrays höher sind, unabhängig von der Boomlänge. Der wirkliche Vorteil der Boxkites ist jedoch deren kürzerer Boom im Vergleich zur Yagi. Auf die G/T-Werte der beiden Antennen werden wir später

nochmals eingehen.

Vergleich von Einband-Yagi und Dualband-Boxkite in gestockter Anordnung

Ein Vergleich von Dualband-Boxkite-Arrays ist schwierig, da es, soweit ich weiß, keine gestockten Dualband-Yagis gibt. In diesem Fall denke ich, ist es fair, Yagi-Arrays mit der gleichen Boomlänge wie die der Boxkites zu vergleichen. Ein Array von EF0206 Yagis ist 2,4m lang mit einem Abstand von 2,6m in der E- bzw. 2,3m in der H-Ebene. Eine vergleichbares Boxkite-Array bestehend aus 5 Antennen mit je 13 Elementen, ausgelegt für 2m und 70cm, ist etwas kürzer als 2 Meter und hat Abstände von 2,8m in der E- und 2,5m in der H-Ebene. Die Gründe, warum wir 5 Boxkites nehmen, können in [1] gefunden werden. Die Ergebnisse für das 2m-Band finden sich in den Bildern 32 bis 35. Die EF hat im Vergleich zur Boxkite eine um 20K niedrigere Antennentemperatur (Bild 37), und der G/T-Wert ist um 0,7 dB besser (Bild 38). Der Gewinn bei 144 MHz ist bei der EF um 0,4 dB höher. Für das 70cm-Band ist die Vergleichs-Yagi eine 4-fach gestockte EF7012. Die Resultate sind in den Bildern 40 bis 42 dargestellt. Das Yagi-Array ist zwar um 11,2K kühler als die Boxkite, der G/T-Wert dagegen ist um 1 dB niedriger. Die Boxkite zeigt auch einen um 2,4 dB höheren Gewinn. Die Ergebnisse zeigen, daß die Boxkite im 2m-Band brauchbare Leistung liefert, aber dem Yagi-Array unterlegen ist. Im 70cm-Band ist es umgekehrt. Hier dominiert die Boxkite über die Yagi-Anordnung. Allerdings hat die Boxkite im 2m-Band zwei Vorteile. Es sind dies das gute SWR und der Gewinn/Bandbreite-Verlauf.

Einfluß der Vorverstärker-Rauschzahl bei praktischen System-G/T-Werten

Wie schon erwähnt, beeinflusst die Vorverstärker-Rauschzahl den G/T-Wert des Gesamtsystems. Bei einer, auch für Sendezwecke benutzten rauscharmen Antenne, haben wir mehrere Rauschquellen: Kabel von der Antenne zum Umschaltrelais, das Umschaltrelais, Kabel vom Umschaltrelais zum Vorverstärker und der Vorverstärker selbst.

Wenn wir idealerweise mal davon ausgehen, daß die Kabel und das Relais keine Verluste haben und der Vorverstärker ausreichend Verstärkung hat, so daß nur er alleine die Empfangssystem-Rauschzahl bestimmt, dann kann man leicht den Einfluß eines möglichst rauscharmen Vorverstärkers auf den Antennensystem-G/T bestimmen.

Wir nehmen die Frequenz von 432 MHz und eine Rauschzahl von 0,4 dB an. Der Rauschfaktor F_n ergibt sich aus

$$F_n = 10^{\frac{NF}{10}}$$

und die Rauschtemperatur T_r

aus

$$T_r = (F_n - 1)290 \text{ K}$$

Für unsere ausgewählte Rauschzahl beträgt somit $T_r = 28\text{K}$. Um nun den daraus resultierenden G/T-Wert für Antenne und Vorverstärker zu berechnen, addieren wir T_r und T_{ant} und dividieren anschließend durch den numerischen Antennengewinn G_n

$$G/T_s = 10 \log_{10} \left[\frac{G_n}{T_r + T_{ant}} \right] \text{ dB}$$

wobei G/T_s der System-G/T und $T_s = T_r + T_{ant}$ ist.

Nun wollen wir den Vergleich zwischen der 21-Element YU7EF und der 24-Element DJ9BV-Yagi wiederholen. Als Referenzwert nehmen wir G/T_s und die Vorverstärker-Rauschtemperatur von 28K. Die Ergebnisse in Bild 43 zeigen einen Unterschied von gerade mal 0,05 dB anstelle von 0,45 dB beim Vergleich der Antennen ohne Vorverstärker (Bild 10). Diese Differenz ist nicht mehr wahrnehmbar. Bei allen Systemen, in denen mehrere Mechanismen die Gesamtleistung bestimmen, bringt es wenig, wenn ein einzelner Parameter positiv herausragt, denn dadurch werden oft andere Parameter negativ beeinflusst, wie in diesem Fall das Antennendiagramm und das SWR. Ein weiterer wichtiger Punkt ist die Tatsache, daß manche Antennenbauer absichtlich den Gewinn reduzieren, um den G/T-Wert zu verbessern. Vom Standpunkt des G/T_s aus gesehen, ist das nicht gut, wie uns ein Beispiel zeigt. Angenommen wir haben eine 70cm-Antenne A ohne ohmsche Verluste und ohne Nebenzipfel und sie "sieht" 15K. Dann entspricht die Antennentemperatur der des Himmels

$$T_{ant} = 15 \text{ K}$$

Der Verstärkungsfaktor G_n sei 150 (21,76 dB) und damit der G/T $150/15 = 10$ (10dB). Als nächstes nehmen wir eine Antenne B mit einem Gewinn von 180 (22,5 dB) aber mit Nebenzipfeln und ohmschen Verlusten und einer Temperatur von 20K. Der G/T-Wert ergibt sich zu $180/20 = 9,0$ (9,5 dB), also 0,5 dB weniger als bei Antenne A. Bei Verwendung eines Vorverstärkers mit 28K Rauschtemperatur ergibt sich:

$$G/T_s (A) = 10 \log_{10} \left[\frac{150}{(28+15)} \right] \text{ dB} = 5,43 \text{ dB}$$

$$G/T_s (B) = 10 \log_{10} \left[\frac{180}{(28+20)} \right] \text{ dB} = 5,74 \text{ dB}$$

Nun ist der effektive G/T-Wert der Antenne B um 0,3 dB besser als der von Antenne A.

Der Vorteil eines reinen Antennenvergleichs ist der, daß eventuelle Verbesserungen in der Verstärkertechnologie Art nicht in das

Ergebnis eingehen. In der Praxis bedeutet dies, daß Antennenbauer möglicherweise falsche Entscheidungen hinsichtlich Designoptimierung treffen. Deshalb denke ich, daß ein reiner Antennenvergleich wenig Aussagekraft hat.

Zum Schluß wollen wir noch G/T_s von Boxkite und Yagi auf 70cm vergleichen. Die Ergebnisse in Bild 44 zeigen, daß bis zu einer Boomlänge von 8λ der G/T_s der Boxkites höher als der jeder Yagi ist. Es zeigt sich auch, daß bei Boomlängen $>6\lambda$ G/T_s der K1FO-Yagis über denen der YU7EF-Yagis liegt. Der Grund ist vermutlich, weil letztere den Gewinn reduziert um einen guten G/T -Wert zu erreichen.

Ich habe hier vergleichende Daten von Boxkite und Yagis präsentiert, die vor allem für EME- und Weltraumapplikationen wichtig sind. Obwohl die Boxkites ursprünglich nicht für solche Anwendungen entwickelt wurden, sehen wir doch, daß sie mit den Parametern moderner Hochleistungs-Yagis sehr gut mithalten können. Besonders die Dualband-Boxkite zeigt trotz der nötigen Kompromisse beim Bau eine unerwartet gute Leistung.

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5. Brian Cake, KF2YN, "Boxkite Yagis, Part 2", QEX, American Radio Relay League, September 2004

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FT-817 Frequency Reader for Transverters

"Let me do your microwave maths"

by Paolo Cravero, IK1ZYW – ik1zyw@yahoo.com

In 2008 I designed an external keypad allowing direct frequency entry for the FT-817(ND). The circuit is powered from the radio, with a microcontroller and a 16-key input user interface. Then in 2010 I was asked to design an external display for use in mobile installations. When completed, I received feedback that it would be nice to make the external display take into account a transverter local oscillator, have it display the microwave operating frequency. It looks like a good percentage of transverters are driven by an FT-817, so I took up the challenge and will introduce you to the result.

Hardware

All versions connect to the CAT port of an FT-817 (the FT-857 and 897 are also supported), so the choice of the microcontroller was heavily influenced by the need for a serial port (USART). I also wanted to be able to use cheap and reprogrammable chips in DIP packages, so that anyone could reproduce my projects. Some Atmel AVR controllers have both these characteristics, like the ATtiny2313 used for the stand-alone keypad and the ATmega48/88/168 series, used here, priced ~1.50€ and ~3€ respectively. Since the Yaesu CAT protocol provides no communication error checking, an exact baud rate must be generated, so a crystal must be used for the uC clock. The datasheet gives a table of possible selections. Using a 11.0592 MHz crystal excludes harmonics falling in the lower ham bands (<UHF). The LCD display is the 2 x 16 de-facto standard using a Hitachi HD44780 chipset. It can be driven using only 6 I/O pins and some models have a backlight as well (16-pin connector), with prices starting from ~10€. Add a 78L05, a few caps and you can build the Frequency Reader (FR) for less than 20€ (in 2010) on perfboard/veroboard.

Software

Since I do not operate on microwaves, I have no experience of the operating practices and equipment standards used today. I exchanged some emails with few hams and learned that 100 Hz frequency stability and accuracy up to 140 GHz and possibly above is quite usual. So showing the transverter output frequency to 100 Hz resolution was a strict design requirement. Another requirement was to have a reprogrammable LO value and to retain it after power off of the "Frequency Reader" (the final name was chosen when development was almost finished). Last but not least it seems that some operators use two transverters at a time with the FT-817, so this has to be somehow supported in my accessory.

Summary of Frequency Reader features:

- 9600 baud CAT rate (Yaesu's default)
- ~300 ms frequency update interval
- LO values ranging from 0 to 214 GHz¹, stored in EEPROM
- LO value accuracy down to 100 Hz [ex.10,224,012,3(00) Hz]
- one or more LO values can be saved
- display of output frequency and RX/TX mode
- "configuration" and "normal" operating modes
- two button user interface
- support for both sum and difference mixing schemes
- automatic FT-817 setup for dual-transverter operations



Operation and Versions

Since several versions of the FR firmware have been developed and tested in my shack, I will not go into details here, but full documentation is available through my webpages. As an example, it is possible to have the FR control a band switching relay for those using two transverters at a time. Or to have the FR select the LO value according to the VFO frequency (10.224 for 144 MHz and another for 432 MHz). Or associate a "band center" frequency with each LO value, so that the '817 automatically QSYs to the proper frequency when selecting a transverter LO. A general-purpose version of the firmware is available for download through my webpages, while customizations can be arranged privately. There are Video demonstrations on my Youtube channel. The FR has two operating modes: "configuration" and "normal".

¹ The maximum displayed frequency, IF + LO, is 214 748 364 700 Hz

With configuration mode it is possible to set the transverter(s) local oscillator(s) using the two button interface. The LO value is stored in the on-chip EEPROM, so it is retained after power-off. Configuration mode is accessed at FR power-up, by pressing the [Up] key while the welcome message is shown. In normal (operating) mode the circuit periodically reads the QRG/mode information from FT-817/859/897 from the CAT port, sums (or subtracts) the selected LO value and displays the microwave transceiver frequency. Depending on the firmware, [Up] and [Enter] buttons have different functions, such as changing the active LO value or swapping VFO A/B. Now, simply tune your transceiver while looking at the Frequency Reader, and you will know where you are on microwaves.

Circuit Diagram and Future Developments

While this article was being written and published, firmware development has continued giving birth to a working (but not extensively tested yet) variation called the "Interactive Frequency Reader" (IFR). This is a combination of the stand-alone keypad and a Frequency Reader. It allows text to be displayed to the operator via the 2x16 LCD display, and with full interaction through the keypad. The IFR will be described in a future article. In order to simplify maintenance and components supplies, both the FR and IFR make use of a common circuit diagram including the microprocessor, with the IFR requiring only the addition of an I/O extender chip, 8 resistors and the matrix keyboard. Interested operators may first build the Frequency Reader and then extend it: just add the extra components and reprogram the ATmega microcontroller. No PCB design is supplied for the time being. Please note that the published circuit diagram shows the IFR components as well. A couple of notes about the display: more detail is in the downloadable manual. The LCD backlight current is not standard across display models and it might exceed the 100mA rating of the 78L05. In any case the CAT port is not designed to supply more current, so measure the current needed and consider powering the backlight from a dedicated 9V battery or from the FT-817 supply. The LCD backlight pin out might be reversed too, so test it before soldering. The LCD contrast control is not standard and some displays require an intermediate voltage between 5V and 0V (on the circuit diagram pin 3 is connected to ground). Check the datasheet or test your display before final installation. A general purpose firmware version is available for download via my webpages. Pre-programmed chips are available at cost. Feel free to suggest improvements to the Frequency Reader.

References

[1] Keypad, FR and IFR pages, including docs and firmware (or click on the main website below or scan the QR-code):

<http://spazioinwind.libero.it/ik1zyw/hardware/keypad/>

[2] Author's main website:

<http://www.paolocravero.tk>

[3] Author's blog, with updates on all projects:

<http://ik1zyw.blogspot.com>

[4] Author's video collection:

www.youtube.com/user/ik1zyw



Fig. 1: Frequency Reader in operation with a test L.O. of 10224000.0 kHz. All the circuit fits under the display. The 10x5cm board also holds the keypad option (see text).

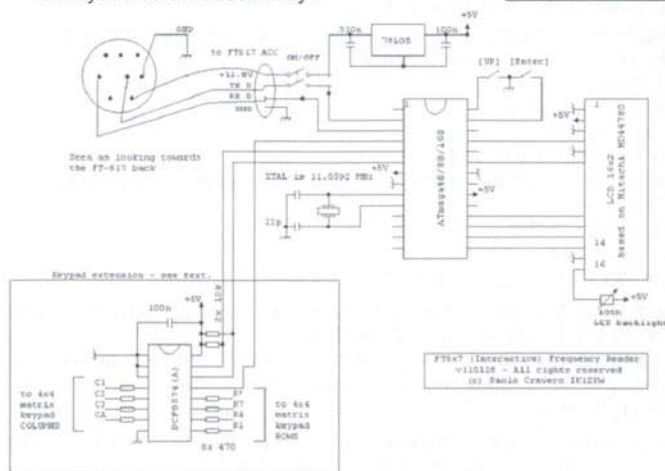


Fig. 2: Combined circuit diagram for the Frequency Reader and Interactive Frequency Reader (see text).

Mid-latitude Sporadic E on 2m - Questions and Speculations

by Angel Nestorov, LZ1AG

My interest in mid-latitude Sporadic E or E_s propagation on 2m began 35 years ago, when I made my first QSO. At that time I became an E_s addict and now I realise that my addiction has become chronic. Between May and August every year, my aim is to be as close as possible to my home and to try to guess when there will be an E_s opening. I have searched for and read many publications in journals and on the internet, and have published several articles myself. Nevertheless, I am aware that I may not have learned everything and certain things may have escaped me. Therefore some of the questions I put below and in particular the answers to them may be the fruits of insufficient knowledge. Yet I would like to express my opinion on certain questions which merit an answer or perhaps provoke an answer from somebody who is more knowledgeable than me.

Why is it possible to make a QSO with QRP during E_s propagation?

Have you ever had a QSO during an E_s opening with received signals of S9+20dB or even +40dB? You will probably answer: Yes, sometimes, but the strength of the signal usually varies and often goes down from S9+20dB to S0 within just a few seconds.

And have you asked your contact what power and what antenna they use in order to generate such a strong signal? You will probably find it incredible if the answer is only 10W and with an HB9CV antenna or similar.

Many years ago, on 10 July 1978, having dismantled my 2m antenna, stations from France and Germany became clearly audible in the 100MHz FM band. My immediate reaction was to use my short-wave long wire and try to adapt the output of my 2m transmitter to it. Fortunately, I was still using homebrew equipment which made it possible to match the antenna. When I finished this, stations could already be heard working on E_s on 2m. To my great satisfaction I made 39 QSOs, with almost all reports S9 (DUBUS, 1, 1979).

Another interesting phenomenon that I have witnessed many times is that shortly before the opening of E_s propagation on 2m, stations on 6m with signals of S9+60dB become audible. I use an ICOM 706 MK2 and a 3 el yagi.

I have tried to calculate what power a station located 1500km away should have in order to receive it at S9, assuming that I receive a direct signal in free space and the two antennas are directed towards each other, each having a gain of 12dB with 2dB feeder loss. This is the result:

- An S9 signal at the antenna input of the receiver should have an amplitude 5μV equivalent to a power 0.5pW at 50Ω impedance (standard IARU), which means $P_2 = -93\text{dbm}$.
- The losses along Tx-Rx chain include:
 - feeder loss of the receiving antenna $L_2 = 2\text{db}$
 - gain of the receiving antenna $G_2 = 12\text{db}$
 - free space path attenuation for 1500km $\gamma = 140\text{db}$
($\gamma = 32.5 + 20\log D + 20\log f$ dB, where D is the distance in km and f is the frequency in MHz)
 - transmitting antenna gain $G_1 = 12\text{db}$
 - transmitting antenna feeder loss in $L_1 = 2\text{db}$
- The output power of the transmitter is then:
 $P_1 = P_2 - \gamma - G_1 - G_2 + L_1 + L_2 = -93 + 140 - 12 - 12 + 2 + 2 = +27\text{ dBm}$

+27dBm is about 0.5W. This is indeed QRP. But if we add another 20dB, the transmitter power would be 50W, and if the signal increases by 40dB – 5kW. But this is no longer QRP!

So, is it possible during an E_s propagation to receive a signal which is stronger than the one propagating directly in space? My intuition tells me that the answer should be: **yes, under certain circumstances!** Let us start with the physical model of an E_s zone.

Perhaps led on by the term layer, most authors who comment on this model accept the E_s zone as a thin layer within the E layer. According to Whitehead, "very thin layers of unusually dense ionization are only

one or two kilometers thick. These dense, thin layers reflect radio waves like a mirror, with greater efficiency and less absorption than the normal E region" (QST October 1997). This creates the impression of something similar to a flat mirror located parallel to the Earth's surface, from which radio waves are reflected at an angle equal to the incident angle. This leads to another notion, namely that the E_s area is located midway between the two stations. According to Mike Willis, G0MJW, "For a signal to propagate via E_s a cloud of sufficient ionisation must exist at the mid-path point" (see <http://www.mike-willis.com/Tutorial/sporadicE.htm>).

The other notion is that within the E region, there are areas (called E_s clouds) with a higher concentration of ions, inside which the concentration of ions continues to increase. The E_s area which can reflect 144MHz is probably inside another E_s area which has a lower concentration of ions, and which is less concentrated, but covers a larger area. That is why E_s propagation on 10m precedes the appearance of E_s propagation on 6m, which in turn precedes propagation on 2m. Consequently, the areas over which it is possible to establish a contact during an E_s propagation are the smallest on 2m and larger on 10m. Unfortunately, I cannot support this statement with data, but my experience over the years leads me to this conclusion. Glenn Hauser writes: "It builds up from low frequencies to a certain maximum usable frequency (MUF) which may vary widely from minute to minute, and opening to opening. E_s always hits the lower frequencies first". (<http://www.anarc.org/wtfd/propagation.htm>).

Each E_s cloud probably has different dimensions, and zones with different densities of ions are positioned within it in different places, but the zone with the highest density of ions takes up an insignificant volume. According to Whitehead, "Some E_s clouds are only 100 meters across and have critical frequencies over 30MHz, yielding MUFs over 150MHz" (QST November 1997).

In such a case, the refraction of radio waves will be different in different zones of the E_s cloud so it will behave in a similar way to a curved mirror. The zone with the highest concentration of ions will act like a concave mirror, which can collect and direct more energy to some points than a direct or refracted path would.

Is the E_s cloud located at the mid-path point between the two stations?

The E_s zone which can reflect 144MHz is probably a very dynamic zone within which there is a constant movement and change in the concentration of ions, as a result of which the zone changes its shape as well as its location in space. This in turn causes changes in the conditions under which radio waves are refracted, and consequently changes in the area illuminated on the surface of the Earth. This is why during an E_s opening, a station can disappear and reappear, but in most cases after it has died away it does not reappear. At the same time, a few kilometers away it may not be audible at all. Unfortunately, there has not yet been a radio amateur experiment which follows the area illuminated by a station during an E_s opening as it changes over time. The aggregation of information from QSOs made within a given period of time suffers from the failing that each QSO is registered only once. This is why if during an E_s opening the contacts of a station are located within a given geographical area, this does not mean that this station was audible to all stations within the whole area throughout the E_s opening. By way of comparison, one may think of a circus arena where a spotlight lights up different parts of the arena. It is likely that the form and dimensions of the E_s cloud, and above all of the area within it where the concentration of ions is sufficiently great to refract the radio waves with frequencies above 144MHz, will define the focus of this "spotlight", the surface area it lights up and its direction.

Here I want to make two important suggestions:

1. During an E_s QSO the cloud is located between two stations, but not necessarily at the mid-path point. Depending on the concentration of ions and hence the MUF, the distance between the E_s cloud and the two stations differs. If for a given MUF the E_s cloud is at the mid-path point, at a lower MUF it will be possible to establish a contact with a station further away, and at a higher MUF with a station who is nearer.

2. Within an E_s cloud, radio waves change the direction of their propagation not only along the vertical plane (downwards), but also along the horizontal plane (sideways). This depends on the specific shape which an E_s cloud takes during a given period of time.

On maps which show radio paths, a straight line is drawn between two stations when the map uses azimuth equidistance projection. Today this is done using computer programmes (DK5YA, NA3T & NV3Z, SM3BSJ), but the underlying idea is not so different from the manual map which I saw published for the

first time in DUBUS 6/7, 1974. These modern maps are very appealing visually, but in my view they are somewhat misleading. The straight line (according to the chosen projection of the map) is the shortest distance between the stations, but it is unlikely to be a true projection of the signal path. This would be the case if the reflection by the E region was like a mirror placed parallel to the surface of the Earth. Then the ionised area would probably be located above the mid-path point between the two stations, and the intersection of the lines of several radio contacts made at the same time would also be located there. But this is hardly ever the case. In most cases, the intersections of the lines of several radio contacts made at the same time are scattered, and the mid-path points between pairs of stations do not coincide with the intersections and are scattered over an even larger area.

A similar situation is observed on maps with cylindrical projection, where a radio contact is represented by a curved line, but the mid-path point and the intersections are treated similarly. (DF5AI, EA6VQ). Things would look different if we could draw the projections of the radio waves' paths as refracted or bent within an environment which changes their direction of propagation. The simplest model of such an environment is a sphere in which the concentration of ions increases from the surface towards the centre. Radio waves which enter such a sphere will refract with different angles in both the vertical and horizontal planes. So the refracted radio waves will illuminate an elliptical area on the surface of the Earth. The size of this area will depend on the volume of the ionised cloud as a whole and on the part of it which has a high enough concentration of ions to reflect 144MHz signals. The energy distribution of the illuminated area will also be a function of the volume of the ionised cloud and the concentration gradients of the ions inside it. With a given combination of these factors in a small region of the illuminated area, it is possible that energy which has undergone different refraction paths will be focused and the intensity of the electromagnetic field in that area will be larger than that arising from the propagation of a direct signal in free space. And because this ideal model of a sphere does not occur in practice, all of the above may happen, but the refraction will also occur in the horizontal plane (sideways bending).

The visual representation of radio paths on maps would be more accurate if they were represented as refracted lines (straight or bended depending on the projection of the map) – one line from the location of one of the stations to the place which is under the E_s area, and another line from that place to the other station. Unfortunately, it is not possible to do this immediately after each E_s QSO, but it can be done for activity over a given short period.

By way of an example, I will discuss the situation on 14 June 2009, when an E_s event occurred on 2m between 1045 and 1205 UT. If you look at the map (<http://www.mmmonvhf.de>), you may have the impression that the E_s cloud was located over a large area above Northern Italy and Slovenia. If you follow carefully the E_s evaluation, you will notice that the lines of the QSOs made gradually move northwards. In order to analyse the situation, I decided to follow the QSOs made within short time periods (5 minutes) with an interval of 10 minutes between periods. I tried to fill in certain gaps (QTH loc) or to avoid obvious mistakes. In a few cases, there was a difference in time of one minute. The maps which I drew are presented in figures 1 to 6. I have marked on each of them the supposed location of the E_s cloud, the interpolation being made approximately between the densest crossing points of the lines of the QSOs.

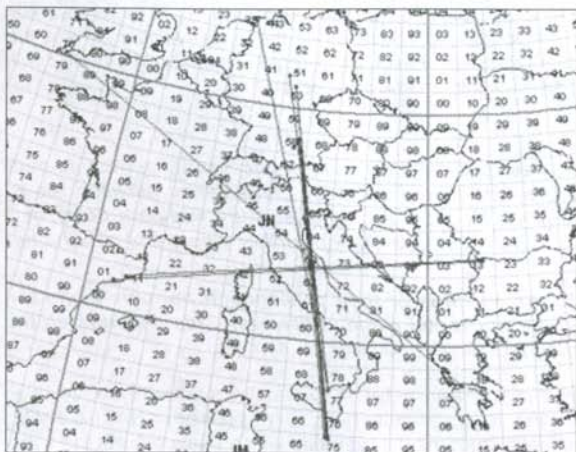


Fig. 1:
1045 – 1049 UT
E_s cloud position JN63LE

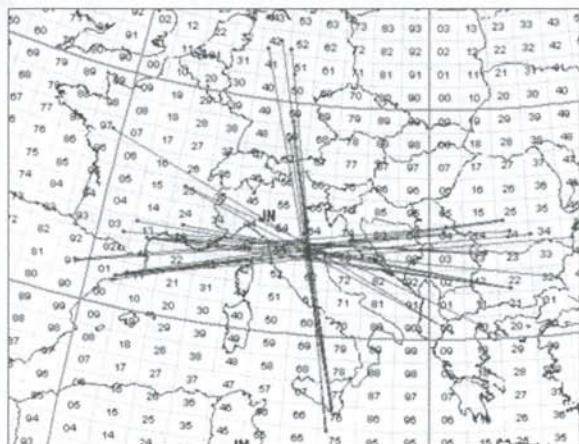


Fig. 2:
1100 – 1104 UT
E_s cloud position
JN63HR

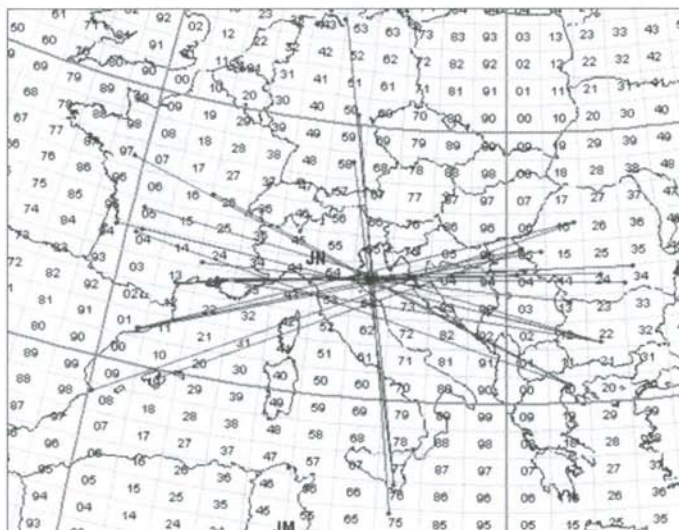


Fig. 3:
1115 – 1119 UT
E_s cloud position
JN64JJ

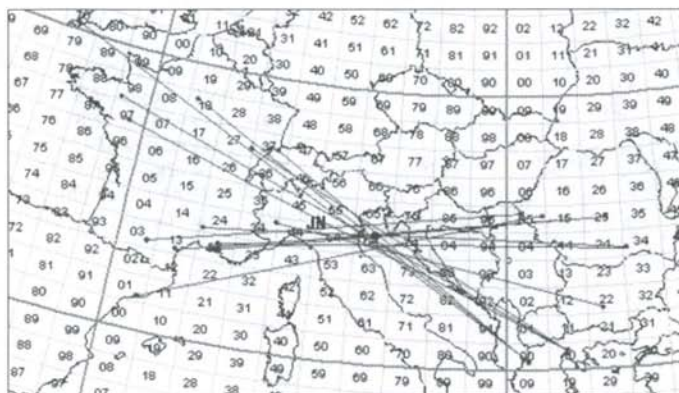


Fig. 4:
1130 – 1134 UT
E_s cloud position
JN64LR

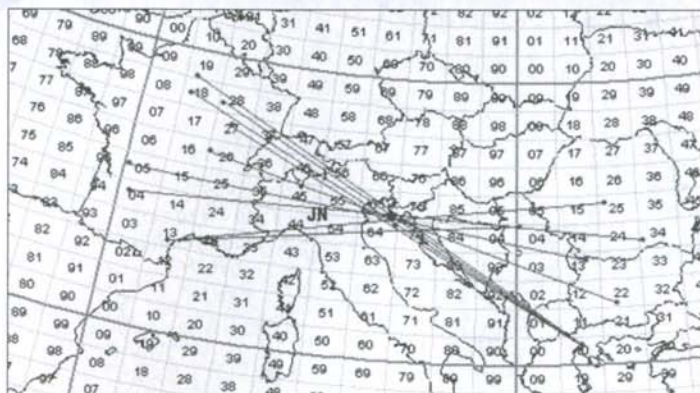


Fig. 5:
1145 – 1149 UT
 E_s cloud position
JN65VC

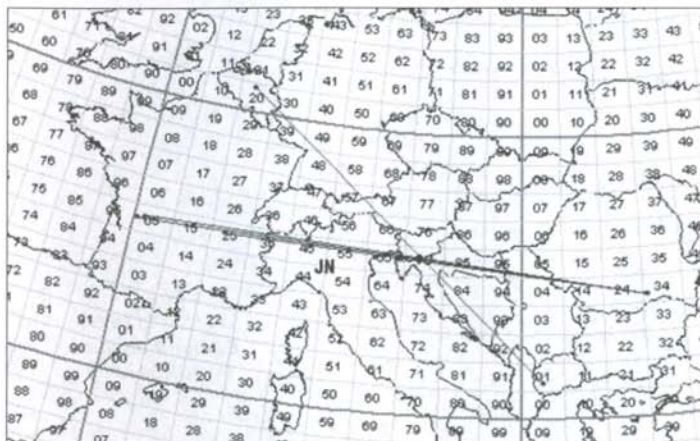


Fig. 6:
1200 – 1205 UT
 E_s cloud position
JN75LN

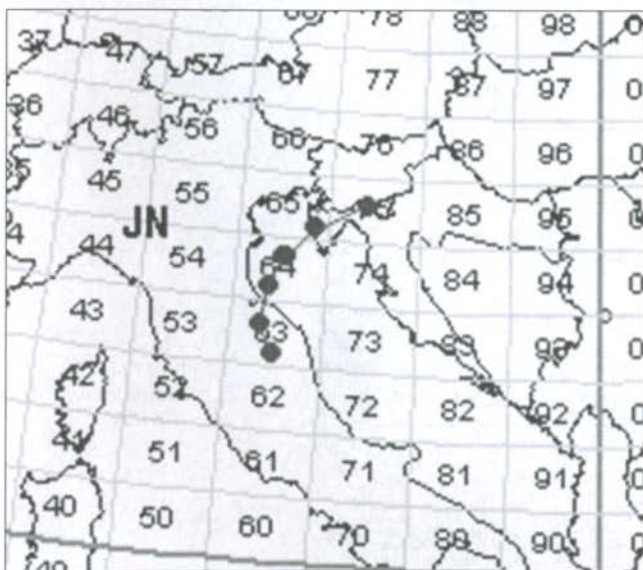


Fig. 7:
Movement of the E_s cloud
on 14 June 2009
in intervals of 15min
1045 – 1049 above JN63LE
1100 – 1104 above JN63HR
1115 – 1119 above JN64JJ
1130 – 1134 above JN64LR
1145 – 1149 above JN65VC
1200 – 1205 above JN74LN

What is interesting about this example?

1. Within each period, the crossing points of the lines of the QSOs are quite well grouped and the location of the E_s cloud is probably correct to within 50km.
 2. For only 2 of the 84 QSOs in this study, the difference in the distances between the two stations and the location of the E_s cloud is less than 10km and it is possible to say that it is at the mid-path point. For 43 QSOs the difference is more than 200km.
 3. Nearest to the cloud was I1DMP from JN34XU at 395km.
 4. The shortest distance worked was between IT9CJC in JM76IW and DL1MJF in JN58OF, at 1278km.
 5. Furthest away from the cloud was G4LOH from IO70JC at 1579km.
 6. The longest distance was between G4LOH in IO70JC and SV3BSF in KM08UF at 2510km.
- Within 75 minutes, the E_s cloud moved in direction SSW-NNE over a distance of some 300km. The direction of movement is different from that indicated in other publications, namely ESE-WNW. (http://en.wikipedia.org/wiki/Radio_propagation#Sporadic-E_propagation).

Which E_s clouds can be used?

In order to make an E_s QSO on 2m, each station's antenna should be able to see the E_s cloud. The upper limit for the distance of an E_s QSO is defined by the curvature of the earth and the height of the E_s cloud above the earth. At a height of some 100km and a distance of 1150km, the cloud disappears below the horizon, which means that double that distance, i.e. 2300km, is the maximum distance between two stations to have mutual visibility of the cloud.

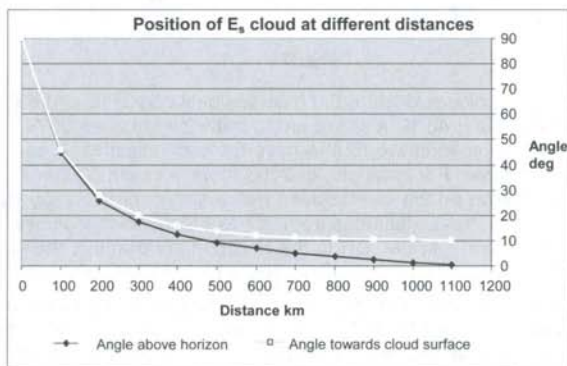


Fig. 8

It is a well-known fact that it is also possible to make EME QSOs when the moon is below the horizon during good tropo conditions. This is probably also true for E_s . At small takeoff angles, the radio waves' path in the dense layers of the atmosphere may reach 200km. The E_s cloud might be somewhat higher than 100km. That is why with a single refraction the maximum distance of E_s QSO may surpass 2300km. For example, on 14 June 2009, the distance between G4LOH (IO70JC) and the E_s cloud was 1579km. At this distance, the E_s cloud was approximately 4 degrees below the horizon. I will not try to guess what caused the bending of the radio waves beyond the horizon. Perhaps another E_s cloud, or maybe some favourable tropospheric conditions? But when one speaks of a record distance for one hop E_s QSOs, distances above 2300km imply the involvement of additional factors.

The lower distance limit for E_s QSOs on 2m is defined by the density of the ions in the E_s cloud and is rarely below 1000km. In Figure 8, which shows the relationship between entry angle and distance, we can see that at distances less than 400km the entry angle increases rapidly. This makes it possible only to use E_s clouds which are further than 400km away. At 800km, the angle at which radio waves enter the E_s cloud is changing very little and this is probably the reason why E_s QSOs are most often made at a distance of 1600 to 2000km. In the sample of 84 QSOs on 14 September 2009, 21 are at a distance of less than 1600km, 51 are within the range of 1600 to 2000km, and 12 are further than 2000km away.

What does mid latitude E_s on 2m mean?

Why is there no E_s propagation on 2m in South America and the Southern parts of Africa?

Why are there more days with E_s propagation on 2m in Europe than in North America?

I have not come across any information on E_s QSOs on 2m made in South America and Africa. I

remember seeing a communication concerning E_s QSOs on 2m between Japan and the Far East of Russia, but I could not find it in order to refer to it. The information I have concerning North America and Australia is rather limited. This is why I tried to compare the latitudes at which E_s clouds occurred during QSOs on 2m made in the period November 2008 to February 2009 in Australia and New Zealand (www.mmmnonvhf.de/es.php) and in the period May - August 2009 in North America and Europe (www.vhfdx.net). I have used the crossing points or the mid-distance point between the stations shown on the QSO maps, or the indication shown in the vhfdx.net software. I am aware of the inaccuracy of this method, but for the purpose of the comparison which I want to make, it is in my view acceptable. The result is shown in Figure 9 below.

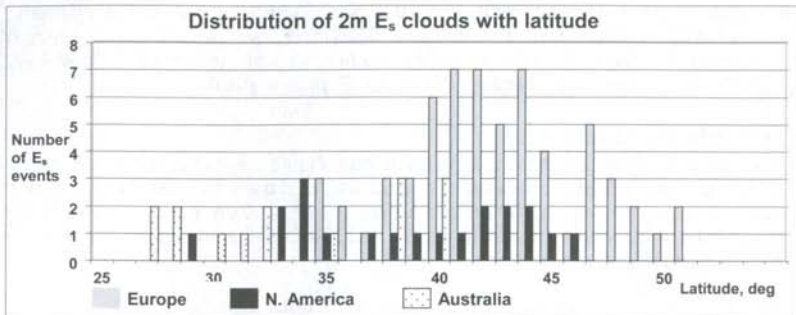


Fig. 9

In Australia and New Zealand, E_s clouds occurred at latitudes between 27°S and 49°S degrees, in North America between latitudes 29°N and 46°N, and in Europe between latitudes 35°N and 51°N. This one-year sample is not sufficiently representative, but is nevertheless indicative. If data for more years are added, these limits will be extended. For example, in 2008 there was an E_s opening in Europe at 58°N. Beyond these limits, E_s propagation on 2m can be used with a single refraction by stations located some 7-8° further away. In Europe, this means latitudes from 27° to 65°N, which includes EA8, CN, EA9, 7X, 3V, 5A, SU, JY, 4X to the South and LA, SM, OH to the North. In my own log, the southernmost QSO on 2m was with 4X1MK KM71GH (31°N), and the northernmost with SM3VAC JP83VA (63°N).

Since there is plenty of information from Europe, I chose at random four years (2002, 2005, 2008 and 2009) and tried to locate the latitude of a total of 279 E_s clouds. I must admit that I omitted a few of them, but they were in the middle of the sample and would not have changed the distribution significantly. On the assumption that my errors have been partially offset by the large sample, I came up with the following diagram (fig. 10) for Europe.

So now I have no doubts about the term mid-latitude for Europe. Most often, E_s clouds occur in Europe at 38-50°N. The outer limits of the four-year sample are 35° and 58°. There may have been a few cases outside these limits in preceding years, but it is unlikely that they were more than 1-2° away.

In Australia and New Zealand, the location of the zone in which E_s clouds may occur is at the lowest latitude. In the sample for 2008-2009, the zone starts at 27°S. Unfortunately, it is not possible to know up to what latitude this zone spreads because at latitudes greater than 40°S, there is no land and events at 40° only occur when an E_s cloud occurs between the southernmost territories of New Zealand and Australia. In North America, the zone in which E_s clouds may occur also starts at a lower latitude as compared with Europe, 29°N, and ends further to the South, at 46°N. I would point out that the year 2009 is not particularly representative and these boundaries should probably be increased by at least 5°.

It is beyond doubt, however, that both in Australia and New Zealand, and in North America, the zones in which E_s clouds may occur are considerably closer to the Equator than in Europe. And vice versa, in Europe E_s clouds may occur further to the North than in North America.

These differences are probably due to the different position of the continents with respect to the magnetic poles of the Earth. It is well known that the magnetic poles are positioned asymmetrically and do not coincide with the geographical poles. Both the Northern and the Southern magnetic poles are located in the Western hemisphere, and the shortest magnetic field line joining them crosses the Hawaii Islands in the Pacific and passes close to New Zealand. The longest magnetic field line joining the two poles (which is some 5,000km longer) crosses Europe. **My hypothesis is that the narrowest zone in which E_s clouds may occur is around the shortest magnetic field line, and respectively the largest zones is**

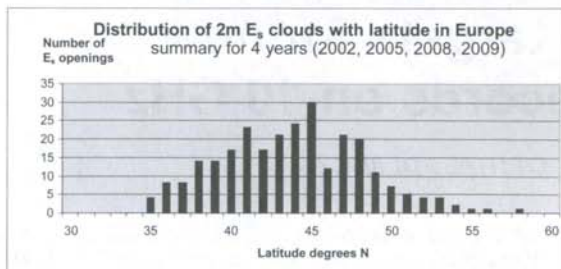


Fig. 10

around the longest magnetic field line. As a result, people in Europe are in a more favourable position! What is more, the combination of the highest latitude with the longest day of the summer is likely to provide one more favourable circumstance for more E_s clouds provoking E_s propagation on 2m, and a greater number of days during this occurs.

Mike Willis, G0MJW, (www.mike-willis.com/Tutorial/sporadicE.htm) has published a map on his website which shows the Magnetic Equator, the Equatorial E_s zone, the mid latitude E_s zones and the auroral E_s zones. I would have limited the zone of mid latitude E_s on 2m on this map, in accordance with what has been said above, to Europe, North America, Australia and New Zealand. It would then become clear that Japan, with the exception of its Northern islands, remains outside this zone, the Southern parts of Africa are far from it, and in the case of South America, only the island of Tierra del Fuego and the Southernmost parts of Chile are within this zone. Some of the Antarctic bases on the South Shetland Islands are at the suitable distance for E_s QSO on 2m, but... you know what the likelihood is of a suitable E_s cloud occurring there. As regards Asiatic Russia, E_s clouds certainly occur there but... there is probably nobody to register them because radio amateur activity on 2m is very dispersed on this vast territory.

Why does E_s propagation on 2m occur mostly during the day?

Why does E_s propagation on 2m occur mostly in the summer?

I put these two questions simultaneously because, in my view, there is a unifying factor in the answers. I assume that the occurrence of E_s propagation on 2m, mainly during the day, is directly linked to solar emissions. But let's not forget that the length of the day changes and at mid latitudes, it is longer in the summer, and beyond the arctic circle the sun does not set during a certain part of the year. What is the situation 100km above the surface of the Earth where E_s clouds are formed? The "arctic circle" there is at approximately 57°N. In the summer, the duration of direct sunshine at 100km above the surface of the Earth in mid latitudes is between 16 and 24 hours. At the same time, it is longer in Europe and shorter in North America and Australia. This fact should not be overlooked.



Fig. 11: For comparison a map covering all 227 2m E_s cluster spots on June 14th, 2009 (Txn to G7RAU / LiveMUF for map!)

Possible targets for new World Records on 10 GHz

by Joachim Kraft, CT1HZE / DL8HCZ

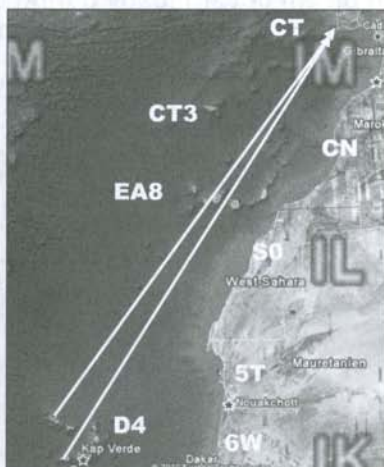
After achieving a new distance world record on 10 GHz over 2696 km between the island of Sal, Cape Verde (D4) and southern Portugal (CT) on 10 July 2010 during the HYPERATLANTICA 2010 expedition, the question is: *What next?*

From the signal strengths and duration of the openings between D4 and CT on 3cm, I would say that at least a bit more is possible on 3cm, and the next distance target could may be the magic mark of 3000km. Also we should keep in mind that N6CA has been trying hard in the past – and especially in the summer of 2010 – to work with Hawaii (KH6) over a 3900km path on 3cm WITHOUT success, although he has made this contact on 5.7 GHz successfully a long time ago. It seems that there is quite a big step between 5.7GHz and 10 GHz, and I think we have seen this also in the signal strengths on the paths between EA8 and CT and D4 and CT in summer 2010. Thus I think that any future world record attempts should not aim for too big an increase in distance.

The question is now where are promising paths for distances greater than 2700km on 10 GHz? Surprisingly there are several interesting options!

1. D4 to CT and EA7 It would be relatively easy propagation-wise to extend the record beyond 2700km with another attempt from Cape Verde. There are several clear paths from more distant islands of the D4 group to EA7 and CT.

Fore example, from Sao Vicente in D4 (HK76MV) to CT (IM57NH) the distance is 2764km. There is already a 2m beacon running from D4C and impressive signals have been heard in CT in August and September 2010. However, this path is a bit disturbed by EA8 and thus may be problematic for 3cm.



Possible paths from other D4 Islands to CT in the 2700 to 2900km range

But there are also other possibilities with bigger distances: From the Island of Fogo (HK75SA) the distance to IM57NH is 2917km and the sea path is totally clear. From HK75SA to IM57OR it is 2959km so we are nearing the 3000km mark. If we go a bit more east to EA7 (Huelva, IM67ME), we reach 3000km and the path is still clear.

I am sure that the probability for such a 3000km QSO in July is quite high. The drawback is the high logistical and financial effort that is necessary for the team on D4.

It is also possible to try again from the same site in D4 (Sal, HK86NU) and on European end of the path to go further north along the CT coast, e.g. to the Lisbon area (IM58JJ looks good). This would give 2740km.

Although there are also options further north in CT and EA1 (e.g. IN52IV, 3169km to HK86NU), I am sure that the probabilities for a 3cm QSO are very low. In the past years even on 2m CT stations from the Lisbon area have almost never worked D4 on Tropo in the summer, although the path from IM57 has been open every day!

2. Across the Mediterranean I remember that HB9RXV, Paul originally wanted to try the path from EA5 to Egypt (SU) but it was impossible for Paul to get licences for Egypt. With this in mind I made a closer tour on Google Earth and found a path from EA5 (IM98QC) to a spot in Israel (KM71GQ) over 3265km. So, this would make a true 'Grande Bleue' event again! But we have to take a closer look on the path as it looks like that is scratching the land in IT9 (IM76MP, Sicily) and 3V (Tunisia). See the following maps. We can see that the path is still free (only over the water) in the IT9 area and the spot in JM76MP would be a perfect location for another 3cm station in between for coordination! Also the path in the 3V area is still free. But this is a very tight path indeed – it is not possible to change the QTH locator in EA5 or 4X by a single letter! If we do this the path is not clear any more, amazing! If there were already teams going to EA5, 4X and IT9, there is an even more crazy spot go for a fourth team: not so obvious from the maps is that the path is also very close to the island of Gavdos (KM24BT, south of Crete, SV9) which is very remote, but still accessible.

The Hepburn propagation maps show many days in any summer where most of this path is orange or red, so there should be fair chances to make a 3cm QSO. The paths from 4X to IG9 and from EA5 to I8 have been worked on 3cm already, so it is "just" a combination of both parts...

The place to go in 4X is called Ashkelon, a city directly at the coast with 100.000 inhabitants, so it should be possible to find the necessary infrastructure. There are even some local radio amateurs who could be contacted, like 4Z5PN Yuri Zinchuk.

On the EA5 side for sure it will be possible to find a QTH at the beach North of Guardamar del Segura.



A 3265km path from EA5 (IM98QC) to Israel (KM71GQ) – is it really clear??



The EA5 to 4X path near southern Sicily, JM76MP (1392km from EA5 and 1873km from 4X)



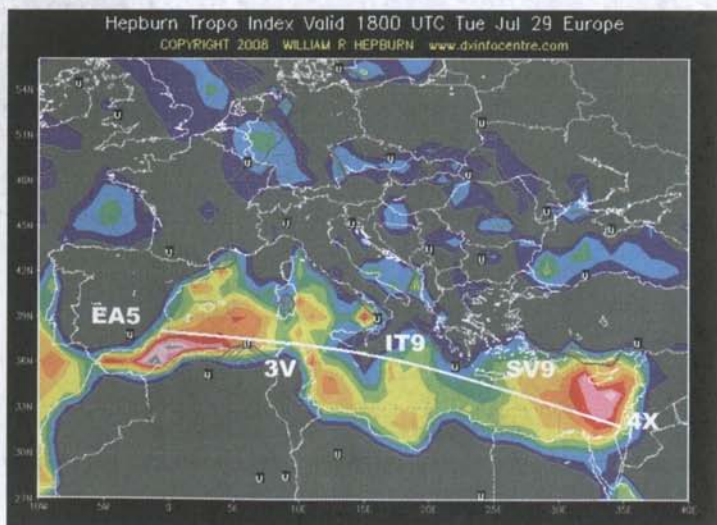
The EA5 to 4X path is very close to the Tunisian coast –but still clear!



The EA5 to 4X path close to the Island of Gavdos, SV9, KM24BT (1029km west of 4X)



The western end of the EA to 4X path north of Guardamar del Segura in EA5 (IM98QC)



The 3265km path from EA5 (IM98QC) to Israel (KM71GQ) on the Hepburn map

3. Italy to Madeira! Another interesting experiment would be a combination of the 'Grand Bleue' and a new HYPERATLANTICA expedition. How could this be possible? What does this mean?

I always like challenging the limits and I have checked a possible path through the Strait of Gibraltar. Although it was very unlikely, I have indeed found a path between Madeira and a spot in southern Italy, that goes clear through the Strait of Gibraltar! This is a bit like winning the jackpot in a lottery.

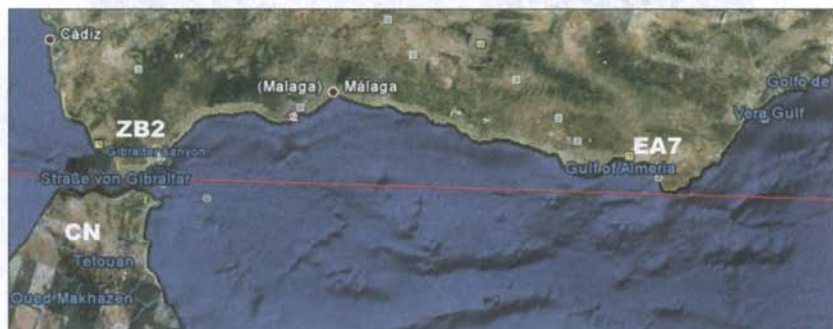
The key locators are: IM12OR on Madeira (CT3) and JM88CV in southern Italy (I8) and the distance is 3036km, so well above our magical 3000km mark.



Is this possible? A free path between Madeira (IM12OR) and Italy (JM88CV)?



Yes! This path is really free inside the Strait of Gibraltar!



The CT3 to I8 path along the south coast of Spain, almost scratching Almería (IM86VR)

We have a close intermediate point east of Almería (EA7, Cabo de Gata, IM86VR) that is 1400km from CT3 and 1640km from I8. A perfect spot for a coordinating team because it is almost in the middle of the path.

The interesting question is of course: Is the Tropo propagation really "travelling" through the Strait of Gibraltar? In September 2010 I discussed with Duarte, CT3HF, about the possible path from Madeira to southern Italy. Duarte was very kind and supplied 2m QSO data from past contacts with stations in "I8". Duarte is quite sure that the QSOs on the maps below were made on Tropo. He reports that there was no 2m Es on that day and no other QSOs were worked at all. In general August 24th would be a late date for 2m Es, although not impossible. We have recorded only one other 2m Es event on August 24th within 38

years in the DUBUS Es data table for Europe. Another important detail at the Italian end is that both of these paths just miss the large volcanic island of Stromboli: one path passes to the north, the other to the south.

Considering all the evidence, it looks like that these two QSOs from CT3 to I8 were genuinely made via sea Tropo ducting through the Strait of Gibraltar.



2m Tropo QSOs from CT3HF (IM12OP) to IW8QQJ (JM88GW, 3069km) and IW8QOT (JM88BR, 3032km) on August 24th 1999



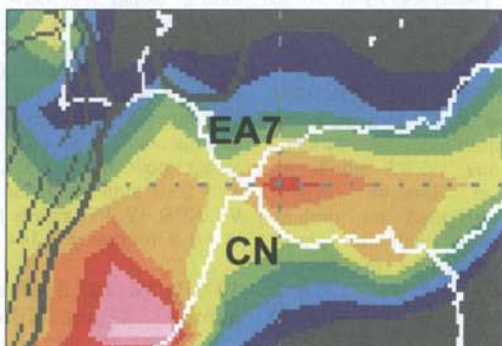
Close-up of the above two QSO paths in the region of the Strait of Gibraltar: The upper one is from CT3HF to IW8QQJ. This was the first QSO at 1406z and 55 reports were exchanged. The bottom line is from CT3HF to IW8QOT at 14:17z where 51 / 54 reports were exchanged.

For a 3cm record attempt, it will again be important to find a free path just north or south of the island of Stromboli. Maybe the best site is half-way between JM88CU and JM88CV, because otherwise the path will come too close to land in EA7 (IM86VR).



The eastern end of the CT3 to I8 path – Stromboli Island has to be considered!

The probabilities for good Tropo paths from Madeira to EA7 and from I8 to EA7 are quite high in July, as we can often see on the Hepburn maps. The remaining question is if the Tropo ducts in the Mediterranean will continue through the Strait of Gibraltar and into the Atlantic, intensely enough to enable a QSO on 3cm with CT3. Hepburn maps sometimes show a break in this area but also often a good probability inside the Strait of Gibraltar itself. For sure it would make sense to have a coordinating and monitoring station in that location at the same time. A perfect spot would be Tarifa (EA7, IM76EA) which is easy to access.



Close up of the Gibraltar area on a Hepburn Map from July 2010 indicates that there may be good Tropo ducting through the Strait



Tarifa (EA7 / IM76EA) would be a perfect spot to monitor and verify the Tropo ducting inside the Strait of Gibraltar



The Tropo needle eye shot from CT3 to I8 seen from space

Final

We can see that there are still interesting opportunities for microwave enthusiasts based in Europe to extend the 3cm world records. Once a 3cm QSO from Hawaii to California has been made, all the above opportunities will be gone – at least, they will no longer be world records. After that there is only the 4000km path from D4 to G left, which is theoretically possible even on 3cm but would need permanent stations on both ends as it is almost impossible to predict when it will open. The teams from HB9 and France which set the latest 3cm world record from D4 to CT in 2010 have decided to try another much longer path on 3cm in 2011 or later, probably from EA8 to the Caribbean (which has not even been crossed on 2m yet...) so the new paths suggested in this article may hopefully be tried by other interested groups! Good luck! 73, Joe, DL8HCZ/CT1HZE

Source and © map pictures: Google Earth

Source and © Hepburn maps: William Hepburn

VHF News

Australia & New Zealand

New 2 metre VK Digital Record

Derek, VK6DZ and Jim, VK3II completed a new 2 metre digital record on 21 January 2011 with JT65b signals up to -1 and -2 dB over a distance of 2497 km – congratulations Derek and Jim.

WSPR – report by Leigh Rainbird VK2KRR

A number of stations have begun to explore the use of K1JT's WSPR (whisper) mode on 144 MHz and more are following suit. WSPR stands for Weak Signal Propagation Reporter. WSPR implements a protocol designed for probing potential propagation paths with low-power transmissions (originally designed for HF). Normal transmissions carry a station's callsign, Maidenhead grid locator, and transmitter power in dBm. The program can decode signals with S/N as low as -28 dB in a 2500 Hz bandwidth. Stations with internet access can automatically upload their reception reports to a central database called WSPRnet, which includes a mapping facility. The WSPR software is available from the WSJT site <http://physics.princeton.edu/pulsar/K1JT/> and the WSPR online database site is located on <http://wsprnet.org/drupal/>

The following stations have been active using WSPR on 144 MHz during January VK2MER, 2KOL, 2KRR, 2XTT, 2EMA, 2QW, 2CDS, 2DVZ, 2BLS, 2DAG, 3SO, 3GHZ, 4LHD, 4FIL, 4VDX, 4JMC, 5GF, 5ZK, 5ACY, 5BC, 5LA, 5AKK, 6DZ and more.

The table below shows a sample of the data you can find online about your signal or reception of others. When on the WSPRnet site you will find it quite interesting to analyse the data and compare 2 or more stations reception of your signal in different locations. There have been numerous surprises for a number of stations finding signal paths they were not expecting to hear, such as the VK5GF to VK2KOL path was very interesting indeed, 1130 Km over the inland and mountains.

The more stations active using WSPR the more interesting the data can become. If you have your 2m SSB rig sitting idle, why not hook it up to WSPR and either let it log reception reports or you can TX

also. Why not have a go? you may be surprised what you might hear or who can hear you, and have a bit of fun.

Note on WSPR Frequency: While the above examples show the use of a dial frequency of 144.144 MHz the Chairman of the WIA's Technical Advisory Committee after consultation has decided that it would be more appropriate to use the international WSPR frequency of 144.489 MHz (dial frequency) in VK. This frequency is within the present beacon allocation but not on the frequency of any existing beacons. Advantages of using the international allocation are that this may open up international 2 metre DX opportunities and would also remove the possibility of mutual interference with SSB stations. Under the guidance of Leigh, VK2KRR, WSPR activity has now been transferred to a dial frequency of 144.489 MHz.

10 GHz Aircraft Scatter using ISCAT and JT65c

David VK3HZ and Rex VK7MO have been exploring the use of WSJT9's ISCAT and JT65c for aircraft scatter on 10 GHz. Both stations use about 60 cm dishes with about 7 watts to the feed and both systems are GPS locked. It has been found that ISCAT works well up to around 650 km and can tolerate Doppler from aircraft crossing the path at more than 30 degrees. JT65c works to near 800 km but requires the aircraft to cross at no more than 10 degrees so that WSJT's AFC can follow the Doppler. The aircraft scatter signals vary widely with some being brief but strong glints of just a second or so that give signal levels of zero to -10 dB and are suitable for ISCAT while others are in the range -20 to -30 dB and of longer duration and suitable for JT65c. It seems likely that the short glints result from specula reflection and that the longer and weaker signals may be the result of minor forward scatter diffraction lobes – the main forward scatter diffraction lobes as used at VHF are just too narrow to get back to the ground and be useful at 10 GHz. Simultaneous experiments are planned to compare aircraft scatter on VHF and 10 GHz over the same path to see if this will provide insights about the scattering mechanism.

Table 1: Sample of Data from WSPR data base
Dr = Drift, P = Power

Timestamp	Call	MHz	SNR	Dr	Grid	P	Reporter	RGrid	km	az
2011-01-22 02:16	VK6DZ	144.145496	-29	0	QF84ux	10	VK5BC/P	PF85mc	1759	95
2011-01-22 00:48	VK6DZ	144.145551	-24	-1	QF84ux	10	VK5BC/P	PF85mc	1759	95
2011-01-21 23:54	VK6DZ	144.145534	+7	-1	QF84ux	10	VK5AKK	PF94ix	1909	96
2011-01-16 22:52	VK6DZ	144.145571	-25	-2	QF84ux	10	VK2KRR	QF34mr	2657	96
2011-01-24 23:08	VK2KOL	144.145479	-12	0	QF56jf	50	VK5GF	PF94hk	1130	256
2011-01-24 22:52	VK2KOL	144.145483	-15	-1	QF56jf	50	VK5GF	PF94hk	1130	256
2011-01-25 03:46	VK5ZK	144.145560	-25	-3	QF56jf	20	VK2EMA	QF37qs	872	68
2011-01-21 07:00	VK3GHZ	144.145402	-15	1	QF56jf	5	VK2KOL	QF56jf	536	34

Activation of QE27 from Mt Owen, Tasmania

Rex VK7MO and Joe VK7JG planned to activate QE27 on the West Coast of Tasmania on 144, 432, 1296 and 10368 MHz on 3 and 4 February 2011. As it turned out Joe had car problems and did not make it. With the reduced manpower Rex did not set up the 432 MHz station. A total of 11 QSOs were made on 10 GHz with four separate groups on the mainland at Mt Gambier, near Geelong, the Dandenong ranges and South Gippsland. Eight of these were on digital and three on SSB. On 144 MHz 17 stations were worked all on SSB. Despite many attempts during the late afternoon no QSOs were completed on 1296 MHz. It is interesting to contemplate why there was propagation on 144 MHz and 10 GHz yet none on 1296 MHz. The answer seems to be as follows:

Initially Rex set up on 10 GHz and worked Colin VK5DK over some 630 km at Mt Gambier on the digital mode JT65c. The Hepburn charts showed there was the possibility of some tropo enhancement. The signal on the waterfall showed no significant spreading consistent with a duct giving the enhancement. Shortly after Rex worked David VK3HZ in the Dandenong ranges and again there was no spreading of the signal.

By midafternoon the South Gippsland group of Ralf VK3WRE, Peter VK3PF and Jim VK3ZYC were operational but there was little evidence of signals - just occasional traces on the waterfall. But some time later the signal strength increased rapidly with wide spreading of the signal. Sufficient to work SSB even though very rapid QSB combined with the spreading made it difficult to copy. This is typical of rain-scatter on 10 GHz.

By late afternoon a further test with the Mt Dandenong group which now included Peter VK3TPR showed signals had faded to just a trace on 10 GHz with no evidence of spreading. At this time Norm VK7AC was reporting strong signals on VHF via a duct from Northern VK7 to VK3 but no sign of Rex's signal on 1296 MHz. The explanation seems to be that the duct was trapping the signal and preventing it getting over the Mountains in central Tasmania. This also seems to explain the initial lack of signals to the South Gippsland group.

A repeat test with VK5DK showed signals had weakened but there was now spreading as evidence of rain scatter.

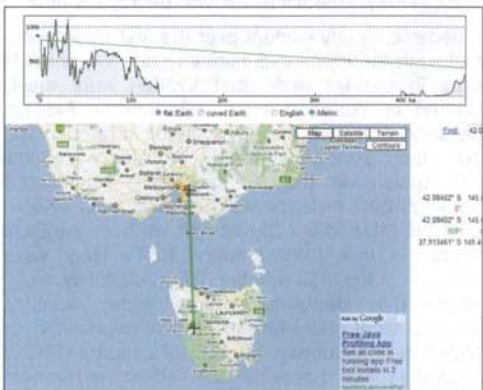
By the early evening the Geelong group led by Chas VK3PY were set up and worked with good signals on 144 but there was no evidence of signals on 1296 or 10 GHz. Rex told them he had to pack up to get off the Mountain before dark but on their pleading agreed to pack up the 144 and 1296 MHz stations and leave the 10 GHz system running. Sometime later the Geelong group telephoned saying they could hear the 10 GHz

signal and an attempt was made at SSB. The SSB signals were distorted with rapid QSB typical of rain-scatter but were too weak to complete a QSO. By going to the digital mode JT65c QSOs were completed with VK3PY and VK3NX. The digital signals showed around 40 Hz of spreading which is an indication of rain-scatter.

Rex believes the most likely answer is that while the duct that developed over Bass Strait in the afternoon had sufficient leakage to allow propagation at 144 Hz it trapped the signals at 1296 MHz and 10 GHz. However rain-scatter which is far more effective at 10 GHz allowed intermittent propagation on this band. Rex has in separate tests from home found that while it is weaker, rain-scatter does also work at 1296 MHz. Thus these initial results suggest that rain-scatter may well be worth exploring as a means of working microwave bands from Southern Tasmania over the mountains in central Tasmania to VK3. (Rex, VK7MO)



VK7MO/p (QE27) beaming to VK3HZ (QF22)



Mt. Owen (QE27) to VK3HZ path

VK9NA to ZL1TPH/p report

To support the VK9NA VHF and microwave DXpedition to Norfolk Island from January the 9th to the 19th 2011, we drove to the top of New Zealand and operated portable out of a vehicle for the same two week period. Two sites were chosen because of easy road access and panoramic elevated views into the ocean looking towards VK9NA. One portable QTH chosen was Ahipara on the west coast and commonly known in New Zealand as the Gum fields, a high 200 m plateau located at RF64NT and 838 km distance to Norfolk Island. The other portable site was at the top of New Zealand, called Cape Reinga at RF65IN and being closer at 748 km to Norfolk Island and at 175 m ASL.

Our prime focus was the microwave bands from 23 cm to 3 cm with 144 MHz as liaison. Our station on VHF was 250 Watts to an 8 element horizontal Yagi on 2 m. With the microwave bands a 1.25 m diameter andrew dish with 150 Watts on 23 cm, 100 Watts on 12 cm, 20 Watts on 9 cm, 100 Watts on 6 cm and 5 Watts only on the 3 cm amateur band. For a numbers of days I was assisted on site by both Simon ZL1SWW and Harry ZL1BK with many thanks on two separate occasions.



ZL1TPH/p RF65IN to VK9NA on 13, 9 & 6cm

It became clearly evident over the first week and a half, with the many excursions to our two portable sites in the far north that VK9NA were easily worked or heard at either 838 km or 748 km distance with their fine 2 m station at around S2, but the conditions only provided marginal propagation on the 23 cm band. However a personal best distance was achieved with VK9NA on 1296 MHz SSB / Digi at 838 km from the Gum fields on the 13th January NZT. They were received at best -12 with the JT65C digital mode. Watching the Hepburn charts as the days went by, we saw what could be an intense inversion layer appearing on Sunday the 16th of January UTC at around 1800 or thereafter, with red on the Hepburn indicative of "Very Intense Propagation" between Cape Reinga (RF65IN) up to Norfolk Island (RG30XW) at 748 km.

Tropical Cyclone Zelia was coming down from the tropics in a Southerly direction towards both Norfolk Island and New Zealand at this time.

Cyclones that come down from the tropics actually diminish in general or become weaker. A high at the same time had moved across the top of New Zealand and was heading Westerly away past Cape Reinga. It was decided with prior emails to VK9NA that we attempt at 1900 UTC on the morning of the 16th January 2011 (8 am Monday 17th NZT)



ZL1TPH/p RF64NT to VK9NA 838km

Driving north from my motel accommodation at 6:30 am in the morning from Kaitia, it was raining and wet and dismal. Arriving at Cape Reinga 1.5 hours later at 8 am, it was surprisingly dry underfoot but in the clouds and mist and the humidity was extremely high with no visibility whatsoever. Initial contact on 2 m was not that strong at only S2 at 748 km distance.

We both decided to set up our respective 1.25 m dishes at each end, and then tested on 12 cm (2.4 GHz) with no success. A call was made for more accurate dish alignment at my end, because with no visibility, I had lost my accurate ridge line bearing at 200 metre distance away, which points to Norfolk Island. We dropped down to the 23 cm and I locked my dish on the VK9NA signal.

With tests again on 12 cm, their beeper identification was easily heard, resulting in a weak SSB contact on 2.4 GHz with headphones on at 10 am NZT. Digital JT65c was then attempted, but the frequency instability at my station resulted in no decode. We moved to the digital mode FSK441 as an attempt to complete a digital QSO on 12 cm which was achieved. Signals at this point were becoming stronger on my receive and VK9NA asked for an additional piece of unknown information sent to me on 2.4 GHz, received was "RECORD 73 RECORD 73 RECORD 73 RECORD" on my PC.

At around 10 am ZL time or 2100 UTC our liaison frequency on the 2 m band surprisingly went up to S9 plus +. At one point, operator Alan VK3XPD reduced power from 400 Watts to 35 Watts on 2 m and was still S9 with no QSB. It turned out we now had a three to four hour intense temperature

inversion. We now moved up to the 5.76 GHz band. The initial received signal from VK9NA was not strong but VK9NA was easily heard at my station at around 11:30 NZT. VK9NA could not hear my return transmission on 6 cm. Alan VK3XPD made an excellent call and asked, "If you can hear us OK, how about going outside and check your dish alignment. I swung the dish only a few degrees to the north and VK9NA were now 57 and extremely loud on 5.76 GHz. We completed on SSB to VK9NA, to operator Kevin VK4UH.

With our 5.76 GHz stations, both were locked either to a GPS 10 MHz reference or a 10 MHz TCXO for frequency stability, we now tested JT65C. Best decode was -5 with callsign exchange along with an additional piece of information sent as an extra requirement from VK9NA, with the words "RECORD STEVE" received at -14.

Once that was complete we attempted a contact on 10 GHz or the 3 cm band. No signals were heard each way, but at the time the TWTA from VK9NA was not fully operational.

A call was then made to move back down the bands and test on 3.4 GHz or the 9 cm band at 1 pm NZT. The VK9NA signal was easily heard on this band also, resulting with a fine SSB contact to station operator, Kevin VK4UH once again. Signal reports from Norfolk Island on this band ranged from 51 to 59 at around 1 pm NZT on SSB. We attempted JT65C on 9 cm but it was soon evident, that my 9 cm transverter was not stable enough for this transmission mode. The line coming down the screen was clear but all over the place. We then changed to FSK441, but shortly afterwards the band started to close. VK9NA was now gone on 9 cm at around 1:30 pm NZT, both ways to our surprise, and so no digital contact resulted.

We checked the 2 m band at around 2:00 pm NZT. The 2m band was down to S2 and the intense opening looked to be over. We tested 10 GHz again with VK9NA now having fixed there 100 Watt TWTA power amplifier, but nothing was heard on 10 GHz from VK9NA or from my station to VK9NA. The intense opening to VK9NA, we believe, only lasted for around three hours that day, from 10 am to around 2 pm NZT on the 17th of January 2011. We both packed up our stations and I left Cape Reinga at 3 pm NZT for another 1.5 hour drive back to the Kaitia motel. The high anticyclone, moving westerly away, across the upper north island of New Zealand, along with tropical cyclone Zelia moving down towards ZL and Norfolk Island we believe created a small and narrow intense inversion layer or duct. The inversion layer itself, which was clearly evident on the Hepburn charts, along with excellent propagation, coincided with an area between two low pressure systems of 970 hPa and 991 hPa. This area will be unstable moist air sinking and spreading out with very high humidity in excess of 90%.

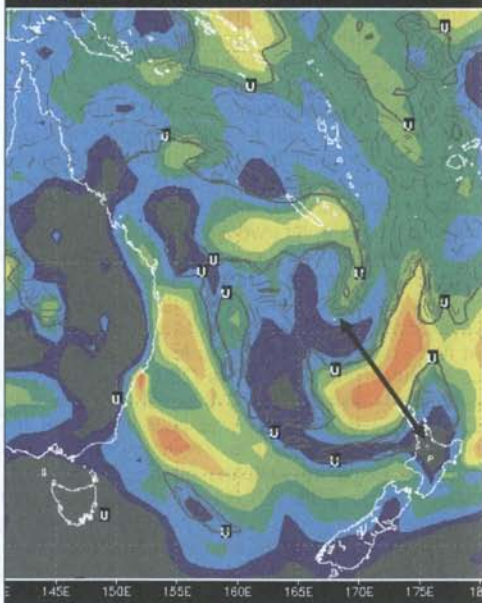


ZL1TPH and his dish with 23/13cm loop feed

Mentioned on the radio, one of the VK9NA operators said, "That the mist was moving in on their operating site at Norfolk Island. They VK9NA were operating at only 50 m above sea level and the Cape Reinga beam heading towards ZL was through Norfolk pines located nearby. Stephen, ZL1TPH

0000 UTC Mon Jan 17 Australia-NZ

M R HEPBURN www.dxinfocentre.com



Hepburn map for January 17th 2011
ZL1TPH to VK9NA on 13, 9, 6cm over 748km

1st VK to ZL QSO on 2.4 GHz

On the 27th January 2011, at 03:26 UTC ZL1TPH worked Adrian VK4OX on 2403.1 MHz SSB with 5/3 exchanged both ways at a path distance of

2314.5 km, from RF73HM in New Zealand to QG63KF north of Brisbane. This is believed to be the first known 2.4 GHz (13 cm band) contact between New Zealand and mainland Australia.

Looking at the Hepburn propagation maps that afternoon, along with the VK4CP VHF logger information, band conditions did look exceptional promising towards VK4 with tropospheric ducting. The VK4CP logger is an invaluable tool and we thank Adam for providing these services to all amateurs.

With seeing Adrian VK4OX residing on the logger, and knowing that he had high performance gear on 2.4 GHz band, I decided to load up my 2 m and 2.4 GHz equipment into my work vehicle, so as to operate portable from a nearby elevated site called Moirs Hill. This site is at 352 metre ASL and has an excellent take off towards VK, and was only a short drive north from my home QTH.

Once on site, I set up my 144 MHz station. A TS700A feeding a 250 Watt SSPA and 8 element 2 m horizontal Yagi and operated from the rear tailgate of my vehicle. The weather conditions were extremely warm, with no winds whatsoever. Within the first few CQ calls on 2 m, worked VK4APG and VK4OX at 5/9 on SSB and this pointed to conditions as being much enhanced.

I and Adrian decided to then test 2.4 GHz. Firstly my 1.15 m dish was set up. This attaches to my 2 m antenna mast which is attached to my toe ball. This dish is fed with the dual mode, 23 cm 13 cm loop feed as described in a recent DUBUS article. From the rear of the dish, a short length of low loss coax is used to feed the 2.4 GHz transverter placed also in the tailgate, beside my TS700A 2 m radio.

With my initial carrier down on 2403.1 MHz, Adrian VK4OX reports it very weakly. This was confirmed by switching off the carrier at my station. My concern at this point, I was not sure exactly where to point my dish. I asked Adrian to provide his CW keyer to me for dish alignment at my end.

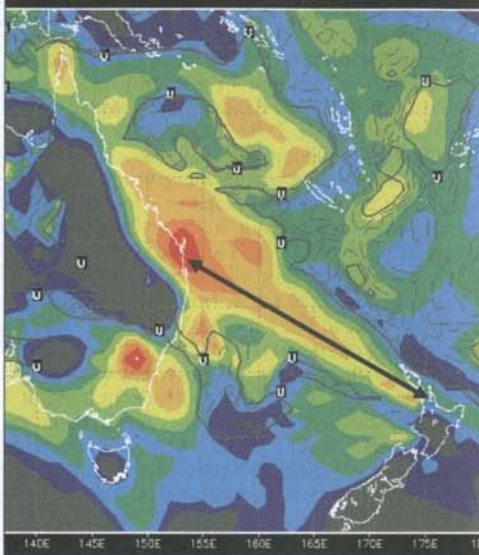
At first I never heard it and a few seconds later it came out of the noise on a QSB peak. I quickly panned my dish and his CW signal was now very strong indeed, my dish was now locked on his signal and I continued to watch my S meter in disbelief. Once he switched off his CW keyer, I called him on SSB and completed the QSO and then chatted.

My 2.4 GHz transverter is mainly home brew, with kit modules mostly from VK. The transverter itself is from Mark VK5EME of Mini-Kits, along with one of his earlier 1 to 2 Watt Pa's and his current 13 cm preamp. TX power is then feed into a VK5KK signal stage Ga-As FET PA, which David produced many years ago. Then into two surplus 75 Watt Spectrian power modules which are combined with W6PQL 13 cm combiners. The 144 MHz IF radio to the transverter at the time was an ICOM 202. I decided not to use my FT817 at the time.

The follow day Brian ZL1AVZ, from RF73FD followed suite with 2.4 GHz contacts to both Adrian VK4OX (2319km) and John VK4JMC (2305km). Stephen Hayman, ZL1TPH

olid 0000 UTC Thu Jan 27 Australia-NZ

WILLIAM R HEPBURN www.dvinfocentre.com



Perfect match: ZL1TPH to VK4OX on 2.4GHz

(tnx to Steve ZL1TPH and Rex VK7MO for submitting the above reports!)

VK9NA - Norfolk Island 2011

The VK9NA team returned to Norfolk Island again in January this year, to attempt to add to the achievements of 2010. With an extra team member, 200kg of gear plus a 2 m EME station, we had high expectations for contacts up to 10GHz. What our best planning could not organise was the weather and the propagation.

In the two weeks we were on Norfolk Island this visit, we experienced 2 Tropical Cyclones, constant high winds and only 3 days of sunshine.

Two metres proved the main activity with 41 contacts by Sporadic E to VK1,2,3 & 7 with a best distance of 2,400, 32 contacts via Troppo to VK2 & 4, and ZL1 & 4 for a best of 1783kms and 15 contacts via Meteor Scatter to VK2 & 4 and ZL3 & 4 out to 1888km.

The bands above 400MHz proved much harder.

70cm produced only 5 SSB Troppo contacts to VK4 to 1502Km.

On Monday 17th January, a 2 hour period of propagation to ZL1 produced contacts on 1.2GHz, 2.4GHz, 3.4GHz & 5.7GHz, both SSB and Digital at 840Km. Then with that, the propagation

disappeared. These contacts will be the basis of 10 application for ZL & VK national records. EME. This year the team added a 2 m EME station based on a 12 metre long Yagi. At least the moon came up each day on time. The EME team worked 124 stations in 96 Grid Squares for the first ever EME operation from Norfolk Island. The hard work was well received with QSL cards arriving every day from all over the world. For more information visit www.vk9na.com Michael Coleman VK3KH



VK9NA 2m EME Yagi



VK9NA shack

VK9NA made the following 25 **first ever** DXCC QSOs on 2m EME in JT65: KB8RQ, JE1TNL, PA2CHR, RK3FG, DK3BU, OH7HXH, EA2AGZ, I2FAK, OZ1LPR, UA9YLU, 9H1PA, UT6UG, SP2OFW, SM7GVF, HA6NQ, F6FAP, EA6VQ, SV8CS, OK1UGA, HB9Q, YO3DDZ, YU7XL, GM6VXB, ON7UC and LZ2HM.

Microwave USA

Editor: Kent Britain, WA5VJB
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Microwave Update 2011

Joint Conference Announcement

Microwave Update 2011 & 37th Eastern VHF/UHF Conference of the Eastern VHF/UHF Society

October 13 – 16, 2011, Crown Plaza Hotel, Enfield, Connecticut- Latest updates, registration forms and hotel info at <http://www.microwaveupdate.org>

Microwave Japan

Seiji Fukushima, Ph.D, JH6RTO
jh6rto@m.ieice.org

New EME station, JH0TOG

Recently one of the Japanese microwave and satellite amateurs, Katsumi, JH0TOG, built his 23cm EME station. He made a septum-fed dish with a diameter of 1.8m to work EME and succeeded in QSOs with HB9Q, PI9CAM, OK1DFC, RW3BP, OE9ERC, 8J1AXA, HB9HAL, HB9MOON, PA3CSG, ES6RQ, and JA6AHB at 10 W. The 1.8-m dish was usually good for most of his terrestrial QSOs although he found it was not sufficient for his 23cm EME operation. Soon after the construction he upgraded it to the present 2.4m dish. Now he outputs 10 W into his dish and uses a 35dB FHX35LGX2 LNA. The output power of 10 W is legal maximum under paper inspection on 23cm. The following is his recent log after upgrading.

All the QSOs were completed on 23cm, JT65.

Date	UTC Callsign	Reports
20 Nov 2010 1705	RD3DA	R-26 R-21
20 Nov 2010 1734	JA6AHB	R-23 R-18
20 Nov 2010 1740	OK2DL	R-21 R-17
20 Nov 2010 1749	PI9CAM	R-15 R-10
22 Nov 2010 1834	UA3PTW	R-27 R-21
22 Nov 2010 1906	OZ6OL	R-20 R-20
18 Dec 2010 1400	YO8BCF	R-26 R-18

In addition he was heard by OE5JFL and VK2JDS so that he expects more QSOs soon. Finally here is his message to DUBUS readers: 'Thank you very much for efforts to work with me while my signal may be extremely weak. When you hear me, please, please try another EME QSO with me. 73



Katsumi's current 2.4m dish

VHF South America

by Flávio Archangelo, PY2ZX
py2zx.ham@gmail.com

Lonh haul contacts between Oceania and South America on 50MHz



Map 1: Not South America, but a very useful map by Bob ZL1RS with the most powerful TV carriers from Oceania, serving as listening guide to find VK/ZL DX opportunities on VHF.
The shift from analog to digital TV in VK is scheduled for 2012.

On January 05, 10 and 12, 2011, Bob Sutton, ZL1RS (RF64vs) and K. Mundell, ZL2TPY were able to copy Jack Henry, OA4TT in beacon mode (Lima, 50,036 MHz 15W) and they made QSOs with him on 50,103 and 50,205 MHz (from Canete FH16tw). Brazilians tried to listen for the ZLs without success. Jack used a homebrew 8 elements optimized yagi and up to 1kW from a converted TL-922a, and Bob running a pair of 6 elements G0KSC and ACOM 1000.

Carl Luetzelschwab, K9LA, tried to understand ZLs QSOs which took place with K6QXY in December 2009 and again 2010 not only taking the Es Multihop theory. His hypothesis considered two Es hops and TEP making the intermediary "chordal hop" function.

More information at:

SSSP: Short-path Summer Solstice Propagation. Han Higasa, JE1BMJ. Published on CQ Ham Radio Japan, September 2006 and on Six News: http://www.ha5hrk.hu/files/SSSP_JE1BMJ.pdf

"Up Over" to "Down Under": W6 to ZL 6meters. Carl Luetzelschwab, K9LA. Published on WorldRadio Online, Part 1 on June 2010 and Part 2 on July 2010:

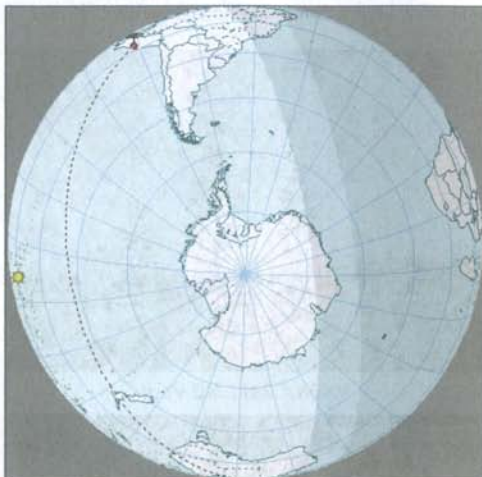
http://www.worldradiomagazine.com/wro_issues/2010/WRO_0710_18-29.pdf

Extreme Multi-Hop 50 MHz Es. Jim Kennedy, KH6/K6MIO. Published 44th Conference of the Central States VHF Society:

<http://www.bobcooper.tv/pdf/Cent-States-2010-Extreme-Multihop-50-MHz-Es.pdf>

ZL/VK websites and forums:

<http://www.qsl.net/zl1rs/> - <http://www.vkloggger.com/>
- <http://www.on4kst.com/chat/start.php>



Map 2: Southern short daylight path between VK/ZL and OA. (Map produced with DX Atlas)

VK4CZ	OA4TT	QG62LP FH16TW CW	2341	2 Dec 09
ZL3ADC	OA4TT	RE66HM 53 10645km	0018	13Dec 09
ZL3JT	OA4TT	RE66GK 419 10647k	0105	13Dec 09
ZL1RS	OA4TT	RF64VS 559 10950k	2341	26Dec 09
VK5ABW	OA4TT	RX 10s then faded	0054	27Dec 09
VK3AUV	CE1/K7CA	QF21WT-FG41 219	11690 km	
			2227	5 Jan 10
ZL4LV	CE3SX	RE54EC-FF46RN 33	9305 km	
			2245	5 Jan 10
ZL2TPY	LU7YS	RF70CV-FE49IU 45	9000 km	
			2300	5 Jan 10
ZL1RS	OA4TT	CW 429 10950 km	2314	23Dec 10
ZL1RS	OA4TT	SSB 51 10950 km	2314	24Dec 10
ZL1RS	OA4TT	CW 529 10950 km	0001	05Jan 11
ZL1RS	OA4TT	JT65a-26dB 10950k	0027	10Jan 11
ZL2TPY	OA4B/B	RF70CV-FH17MW CW	499	10744 km
			2121	12Jan 11
ZL2TPY	CE MUZAK	RF70CV-FF46 WFM	57	9539 km
			0013	14Jan 11
			on 48.3 MHz	

Table 1: Some Oceania – South American 6m QSOs reported in 2009, 2010 and 2011 on DX forums, DX cluster and by ZL1RS.

BC Sporadic-E Reports

Francisco Junior from Itapeperica da Serra, São Paulo (GG66), Southeast Brazil, noted strong signals on Canal 3, LT83 television from Rosário, Argentina on Ch A3 VHF on Jan. 22, 2011 at 2230 GMT. He used only the Prismatic antenna (20 elements) and Semp Toshiba TV. The distance covered was 1.700 km.



Photo 1: Picture of Canal 3 Rosário (Argentina) as captured in São Paulo, Brazil.

Videos at <http://www.youtube.com/franradiotv>

On Jan. 29, 2011 Adriano Rossoni from Rondonópolis, Mato Grosso (GH23qm), Central-West Brazil, captured more than 18 far FM-VHF stations. He described as: "by luck I could listen to one of the best Es apertures that I ever found. At 16 hours Brasília time, I turned on the radio and then they were there with astonishing quantity and intensity. I heard Argentina, Paraguay, [the Brazilian states of] Bahia, Pernambuco, Ceará, Amapá, Pará, Amazonas, etc". The longest distance was 2.400 km and the highest frequency was 95,1 MHz. He used only a vertical antenna PQFM1000, Pioneer DEH P7880MP and Tecsun PL600. Some of his videos are available at: <http://www.youtube.com/adryan0>

93,3 30/01 19:51 R. Amapá FM - Amapá - Macapá - 2.000 km
93,7 30/01 19:52 R. Cultura - Belém - Para - Brasil - 1.800 km
94,3 30/01 20:30 R. FM do Povo - Manaus - Amazonas - Brasil - 1.600 km
94,3 30/01 19:35 R. Metropolitana - Belém - Para - Brasil - 1.800 km
95,1 30/01 20:36 R. Transamérica Hits - Manaus - Amazonas - Brasil - 1.600 km

Table 2 - Partial Es log of Adriano Rossoni

144 MHz TEP Reports

Bruno, PY2NMS (Marília, GG66) made a TEP QSO with Jean FM8DY (FK94mo) on Jan. 30 and Feb. 04, 2011 at 0222 Z on 144,500 MHz FM simplex. He was using IC-2720, 50 W, vertical omni Voyager UVS 300 (MA-6000), while Jean worked with 100 W, Yagi 6 elements horizontal polarization. The distance covered was 4.470 km. FM8DY turns to be the Caribbean reference for 2 meter direct QSOs, as you see on latest spots. Bruno flags a new Brazilian town able to join TEP on higher frequencies. A detailed report of the 2m TEP observations from FM8DY is printed on page 121 & 122 in this issue.

Microwave Europe

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Microwaves UK

UKAC rule changes for 2011

Several changes were made to the RSGB UK Activity Contest (UKAC) Committee contest rules for the 2011 Microwave Bands sections. Previously the only Microwave UKAC event was held on the Third Tuesday to coincide with the Nordic Activity Contest (NAC). However, there were several important differences to the NAC rules. Possibly the most important of these was the contest times. Comments during Contest Forums suggested that the NAC was probably too long and that its start and stop times did not suit UK amateurs, probably because of the 1 hour time difference from much of mainland Europe. The UK event ran (and still does) from 20:00 to 22:30 UK local time.

The second difference was that the same event originally incorporated both 23 and 13cm. However, it became clear that the levels of activity on 23cm were having a negative effect of 13cm activity. Although the two events were scored separately, the interest in 13cm seemed minimal whilst 23cm was a run-away success with ever increasing levels of interest that made it harder to work all available stations within the 2.5 hour contest period. I have already mentioned that the obvious change to making the contest longer was not popular. Many UK amateurs remain convinced

MHZ - DD/MM - UTC - Info - Distance
88,7 30/01 18:35 Unid (tent. Provincia de Córdoba - 1.900 km)
89,7 30/01 18:53 R. D-Rock FM - Mar Del Plata - Argentina - 2.400 km
89,7 30/01 20:24 R. Tiradentes FM - Manaus - Amazonas - Brasil - 1.600 km
89,9 30/01 20:33 R. Comunidade FM - Cametá - Pará - Brasil - 1.700 km
90,1 30/01 20:20 R. Ocidental FM - Rio Branco - Acre - Brasil - 1.600 km
90,3 30/01 18:45 R. FM Selección - Provincia de Córdoba - Argentina - 1.900 km
90,5 30/01 19:03 R. Liberal FM - Unid QTH - Brasil
91,3 30/01 20:25 R. Nazaré FM - Belém - Pará - Brasil - 1.800 km
91,5 30/01 20:05 R. CBN Iranduba - Amazonas - Brasil - 1.600 km
91,9 30/01 20:15 R. Boa Novas FM - Belém - Pará - 1.800 km
92,3 30/01 19:28 Unid (tent. Fortaleza - Ceará - Brasil - 2.250 km)
92,5 30/01 18:55 Unid (tent. Provincia de Córdoba - 1.900 km)
93,1 30/01 19:48 R. Litoral FM - Recife - Pernambuco - Brasil - 2.400 km

that 2.5 hours is about right and that time constraints have generally improved operating skills with far less waffle than in the early days! Since the start of 2011 the UKAC 13cm event has been moved to the fourth Tuesday and now incorporates the higher bands to 24GHz. I suspect that 13cm will prove the most popular band, with 10GHz close behind. It should not be too big a challenge to work everyone on more than one band in 2.5 hours, although if 13cm follows the example of 23cm, it may not be too long before it is necessary to re-plan the contest.

From a personal point of view I find it strange that the RSGB Contest Committee chose to introduce a rule that only UK locators count for score multipliers. With the two contest Tuesdays more or less coinciding with the NAC and other regional contests, introducing a rule that makes working continental/non UK locators less desirable seems to me to be a retrograde step. Whilst the pt/km is still a good incentive to work plenty of continental stations, it ignores the fact that many 'contest' entrants are on to boost their locator squares tally and in doing so contribute to the overall success of the event with their activity.

The first of the UKAC 13cm and up contests, held on the 25th January, saw a useful level of activity and an increased level of interest in 13cm equipment in order to get onto this band. One respondent suggested that in his opinion interest in the bands above 13cm was not as great as might have been hoped. It is early days and the signs are good. Please don't forget there is activity this side of the North Sea and contrary to popular belief (heard at Heelweg) many of us do beam towards the Continental land mass in spite of scoring obstacles placed in our way!

New UK Lightwave distance record

Using red light, high intensity, LEDs G0EWN (with G3PHO) worked G8AGN over a 87km LOS path on the 8th January 2011. This supersedes the previous UK record of 76km set in 2003 by G8LDR/P and G0MRF/P using red light lasers.

The new record was set using baseband modulated light (rather than modulated carrier) and was the result of a slow build up in distance, over many weeks, until the record was achieved late in the afternoon of the 8th January. Reports of 52 were exchanged but it wasn't completely dark and the high gain optical receivers were probably not achieving their highest sensitivity.



Photo 1 shows the red light from G8AGN/P received at Pocklington (IO93PW37) by G0EWN/P at Roper Hill (IO93EI).

G3PHO had his optical receiver with him and although he didn't get to share the two-way distance record, Peter was pleased to be able to give Barry a similar report to that given by Gordon, G0EWN.

Martlesham 2011

Just a reminder that the Martlesham Microwave Round Table will be held on Sunday **17th April** this year. An exciting and informative programme has been prepared and will include talks by W1GHZ, DL4MEA, WW2R, G7OCD, G8AGN/G0EWN and G4SWX and there will be the usual high quality Flea Market and the measurements room. There will be a get together at the Elizabeth Copdock Hotel on the Saturday afternoon and the now traditional Microwave Dinner on Saturday evening. We would love to see more visitors from overseas to this, the premier UK Microwave meeting. Details of the MMRT event are at <http://mmrt.homedns.org/> 73 de Sam, G4DDK -



There are some indications that the former Goonhilly satellite earth station (Cornwall, IO70) will be reopened for educational purposes including EME activities. There are several big dishes available. See picture above.

Calculators on the web

Rogers has a nice microstrip calculator free at www.rogerscorp.com/acm/downloads/mwii/index.aspx Another bunch of microwave calculators is at: www.microwaves101.com/content/calculators.cfm



Heelweg Meeting 2011 – a success with over 200 visitors! Photo by Wolfgang, DD8BD

EME News

70cm & up

Editor: Bernd Wilde, DL7APV

Intro

Dates for **ARI contest** have changed and **ARRL contest** dates are available:

ARI DIG contest 16/17 April

ARI-EME meeting 31 April/1 Mai

ARI CW/SSB contest 24/25 Sept.

ARRL EME contest uW Sept. 24/25

ARRL EME contest 6m-23cm I Oct. 22/23

ARRL EME contest 6m-23cm II Nov. 19/20

All dates are at nearly zero declination what means shorter operating hours for the northern hemisphere, but much more fair for the OPs on the southern part. All 3 weekends are close to perigee and at low background temperatures, so we will see big signals hopefully.

The certificates for the 2009!!! ARRL EME contest were posted last days.

See the moon calendar on my homepage for updates and changes: www.mydarc.de/DL7APV

Just a reminder the **EME NET SAT/SUN 15:00 UT** on 14.345 is still alive. After K1RQG stopped it (thanks Joe for operating the net for so long time) Steve N4PZ took over and runs it every weekend.

In December and January activity dropped a little, but several expeditions kept the PA warm. **CondX** in Jan were partially excellent at perigee with booming signals.

Amateur ham operators have been asked by **NASA** to listen for the beacon signal on 437.270 MHz of the NanoSail-D satellite. It is in a low orbit. Information on reception should be sent to the NanoSail-D dashboard at:

<http://nanosaild.engr.scu.edu/dashboard.htm>.

Expeditions

DL1YMK: Gents, glad to announce that the M & M team will be on the road (?) again for another multiband moonbounce DXpedition from May 28th to June 5th 2011. The activity will cover the 23cm DUBUS contest, as in previous years. The destination will not be disclosed in advance, but it will be an interesting location with a hard to get license. So keep this time frame free from other commitments, which might keep you from giving us a shout off the rock. 73 de M & M

ZA1EME (by OK1DFC ec) is still planned for May this year. (432 and up)

The famous Dutch DX-Team René PE1L, Eltje PA3CEE and Johan PE9DX were qrv as **C56EME** in February on 432 with 50W and one yagi with pol. control. Later in April they will be qrv as **PJ4X** from the new DXCC Bonaire.

BD5RV Michael reports that a group from BY will activate **BY8** (OL14) in mid March with a 11m dish on 70cm. That will produce a huge signal. Later on they will be qrv with a smaller rig from **NL97** on 2m & 70cm and maybe 23cm.



The 11m dish that will be activated from China (BY8) in March 2011

FK/DL2NUD & FK8DD Hermann was delayed by bad wx and problems with local customs. So the YJ trip was canceled. He had several qsos from RG37 and RG28 on 70cm & 23cm and many on 2m.

UN/DL9LBH has got his licence and is waiting for the antennas which will be delivered in spring. Grid is MO51. He will be on 2m & 70cm EME.

EA8/G4RGK: Dave was in the Canary Islands again while on holiday and was able to add VK3UM on 432 CW during a very short window as the moon came up from behind a volcano.

Next 70cm CW ATP dates

March 5/6	Sun 0800-1000 & 1500-1700 UT
May 1	Sun 0430-0730 & 1230-1530 UT
June 26	Sun 0100-0400 & 0900-1200 UT
July 23	Sat 2300-0200 &
July 24	Sun 0800-1100 UT

432 MHz

Several new calls were logged in the last 3 months:

November

I7FNW 4x25 el., 200w
G4FUF 4x39, 130W
K6CLS 1y, 100w
plus US5WU & JN7GVY

December

N9HF 4x22, 50w
plus G4EZF & IK1ODO

January

DJ5BV 2x17el., 600w
VK3GHZ 2x28el., 120W
K6MYC 4y, QRO
WC4N, AJ4MW, VE1MR

KC4KK EN71 is trying to get on 70 cm EME and has 2 and 3/4 of K1FO yagis for 70 cm. They have a polarization rotation mechanism. He is looking for information and assistance in using a pair to get on EME for first time. You can reach Brian at bjrehm@frontier.com.

OK1KIR wkd during Oct and Nov 2010 13 new initials on JT65B: DL2NUD -13/-19 #38; UA3PTW -5/-5; CE0/DK2ZF -22/-24 as 1st CE0 - OK 70cm QSO; PA3DZL -10/-17; WA4NJP -15/O; K4EME -19/O; W7MEM -21/O; K2UYH -10/-9; UT6UG -14/-17; K7QX -12/-18; ES6RQ -16/O and CE/DK2ZF -25/O #49.



PA3DZL SSPA 1296 8xXRF286S water cooled

1296 MHz & up

23cm QSOs worked by **PA3DZL**: 24/10/2010 PY2BS CW&JT, JT: G4CCH, JA1WQF, JA6AHB, PE1HNG. 30/10/2010 CW: OK2DL, SM4IVE, G4CCH, IZ1BPN, F5SE/P, SV1BTR, PA7JB JT65, DL6SH JT65, W3HMS JT65, PA3FXB JT65, PY2BS JT65, YO8BCF JT65, RA3AQ CW, DL0SHF CW, LZ2US CW, N4PZ CW, G3LTF CW, OZ4MM CW, 31/10/2010 OK1DFC CW, JA1WQF JT65, UN6PD JT65, UR5LX JT65, SD3F CW, WA6PY CW, W6YX CW, K1JT CW, VE6TA CW, K7XQ JT65, 02/11/2010 F5VHX CW #121. 73s Jac, PA3DZL

ES5PC reports about 2010 ARRL EME Contest. I was QRV in all 3 weekends, completed QSOs on all **7 bands** where I'm currently QRV. It's one band more than last year (3.4 GHz). Worked both CW and digital as last year. Was there any other station active in this year's contest on 7 or more bands? No station worked on 7 or 6 bands. Only K1JT and W5LUA was worked five bands, a few more on 4 bands. I decided to build a 432 MHz dual-polarization patch feed for my 4.5 m dish. I got the feed ready just before the second 50-1296 weekend and worked 20 stations on 432. It seems this feed is working fine and has a comparable if not better performance than my 4x9w array had. My overall result was down from my last year result (204 vs 221 QSOs, 122 vs 132 multipliers). Number of digital QSOs was slightly higher this year (80 vs 71).

Regarding the microwave part I agree with many others about too many band changes and problems with coordinating the activity between several bands to maximize the number of QSOs. I think it was clearly the largest number of band changes for a single weekend I have ever done so far in my EME history! I wkd:

144 MHz: 53 QSOs, 4CW/49JT, 30 Mult.
432 MHz: 20 QSOs, 11CW/9JT, 15 Mult.
1296 MHz: 68 QSOs, 48CW/20JT, 34 Mult.
23xx MHz: 40 QSOs, 38CW, 2JT, 23 Mult.
3400 MHz: 7 QSOs, all CW, 6 Mult.
5760 MHz: 9 QSOs, all CW, 8 Mult.
10368 MHz: 7 QSOs, all CW, 6 Mult.

My setup on different bands:

144 1kW	4 x 5w M2 H-pol
432 1kW	4.5m dish H/V-Pol
1296 300W	2304 500W 3400 80W
5760 15W	10368 20W all with the 4.5m dish

Many thanks for all the activity during the contest and 73! de Viljo

JA6AHB: PY1KK Bruce and I succeeded on Nov 28th 2010 in a QSO on 432MHz WSJT and CW mode. Distance is 18747 km. Reports were -16/-17 and 559/559, respectively. 18747 km is a

Digital EME distance world record on 70cm. Going back on 26th June 2009 PY2BS and I have succeeded in a CW-QSO on 23cm. The distance was 18853 km and reports were O/O. This should be a 1296 MHz CW EME world record. On Oct 18th 2010 I worked CE0Y/DK2ZF (Rolf, Easter Island expedition) on 70cm in JT65. Reports were -25/-20 On 28th November 2010 I worked BV2A (Bodo, DL3OCH, Taipei expedition 59el. 90w) on 23cm in JT65. Reports were -24/-26. 73 Toshio.

OK1KIR wkd on **23cm** in Oct and Nov 2010: 2nd part of ARRL EME competition – trial operation of new SSPA 1kW. Some heat dissipation problems with existing air forced cooling from tube PA limited our operation to seeking for new stations only. However we wkd on CW on 31.10. F5SE/p 569/539 #305. On JT65C wkd on 30.10. SM0ERR -19/-19 #73; 23:21 UT2EG -14/O and at 23:28 UR5LX -14/O. On 31.10. PA7JB -17/-9; 08:18 UT3LL -24/O; 09:32 UA4API -23/O and at 09:42 SV1CAL -25/O #79. During November we completely redesigned the cooling system of the 23cm SSPA. Now we can use full rf power at CW and JT65 as well. With linear rotate-able dual-dipole feed we wkd in JT65C on 26.11. Bodo as BV/DL3OCH -28/-26 #80; PA0PLY -16/-14; JA6AHB -27/O and G5WQ -27/O. On Sat 27.11. OY3JE -26/-22 #81 and 1st OY/OK 23cm QSO. We found space polarization angle precisely as predicted by F1EHN SW at +75° for BV/DL3OCH signal and -15 for OY3JE! Later on we changed to CP feed and wkd W7IUV -17/-10 #82; LA9NEA -26/O; G4DZU -14/-6; LU8ENU -16/O #83; PA3FXB -12/-7. Afterwards we changed once more again back to LP feed for Bodo's new locator at Taiwan. We wkd at UN7GK -14/O; UY2QQ -22/-18 #84 and finally BV2A -22/-21 #85 again signal polarization turned by +75. The later change back again to CP brought on 28.11. OY3JE -25/-21; IK5QLO -14/-12; PA7JB -15/-12 W5LUA -8/-6 #86. On CW with CP feed we wkd on 27.11. K2UYH 569/569; SM4DHN 579/579, OK2ULQ O/O #306 (dish 3.1m & 180W); SM4IVE 579/579 and G3LTF 569/569. On Sunday 28.11. PA7JB 549/579 #307; I5MPK 569/579, RA3AUB 559/569 #308.

Dec 12 was EME weekend on **13 and 9cm** at **OK1KIR**. Regardless of Moon being close to its apogee and unpleasant windy and freezing rain on our side, there was remarkable MW activity on EME mainly initiated by CT1DMK on 13cm and on 9cm by SM4DHN and S59DCD. Thanks to Luis, CT1DMK we are happy for the first OK-CT 13cm QSO on SSB & CW and new DXCC on 9cm with Lars, SM4DHN. Hrd was Viljo, ES5PC but he suddenly disappeared due to TX failure. In summary we wkd on Dec 12 on 9cm S59DCD 549/559 (=S57NML, so sri no new #), SM4DHN 569/559 as #35 & new DXCC. On 2320 MHz wkd

G3LTF 559/569, CT1DMK 44/54 (sri for RX setting trouble on our side) as #104, IN field and even 1st CT-OK QSO on 13 cm followed in CW CT1DMK 559/559. Tnx also to Peter, G3LTF for a friend's help! 73, Tonda & Vlada for OK1KIR

CT1DMK: I made the very first QSO ever made from CT land on 13cm. Worked DL4MEA with 55/56 straight on SSB upon my call. Both my echoes and Guenter's signals were loud and clear near armchair copy. So all seems to be working just fine, got about 0.4-0.5dB of moon noise although I have a bit of lack of gain in the RX chain that is making the the total system NF of about 0.5dB somehow higher than the preamplifier NF which is 0.32dB. So there are still things to be improved. Power is 150W at the feed (from a 150W amp) as in fact all stuff is at the feed point and 144MHz IF running down to the shack.

Following two weeks I made a total of 22 QSOs, 11 of them in SSB (5 were straight on SSB, so no CW before SSB). #13 initials and presumably a distance record for the 13cm band that is VK3NX-CT1DMK (quite hard to do we have only a few minutes of common moon window). #s are: DL4MEA, SV3AAF, ES5PC, G3LTF, OK1KIR, VK3NX, F2TU, WD5AGO, VE6TA, POBAT, W5LUA, LA9NEA, K7QX. 73s Luis Cupido

Technical

Today I would continue with the technical hints and show some examples with the **VK3UM calculator** what is possible.

For all examples we take average between apogee and perigee and 20°C. The simulation shows the optimum what is possible, in practice you have to add qrm from neighbours, Faraday 0-20dB loss and libration up to 10dB loss!!!

CW the QSO with smallest setups so far I knew was I5TDJ to EA3DXU, 2 yagi to 1yagi.

Gain 19,55 / 16,7 dBD

LNA loss 0,15dB (short cable)

LNA NF 0,35 LNA gain 33 dB

Coax loss 2db RX NF 1,0

Side lobes 10K Back lobe 40K

RX BW 50Hz (good CW operators!)

TX 1500W TX loss 2dB

That gives 0,5dB dB S/N on both sides!!! At perigee 1,5dB, so a qso is possible as the reality showed.

JT65, what is possible with normal average no deep search means at -28dB.

If we take the same equipment as above we need only 100W to run a qso at -28 on both sides. BUT that needs absolut stability on both TRX and no libration. The average what I have seen on 432 in

JT qso's W=2-3. That will drop signal 6dB. Means we need 400w on both sides to complete a qso.

Can OK1TEH with his single yagi work himself? Yes, at W=1 the signals are at -28 and at perigee -27. With a very stable signal and low libration it should work.

Gain 17dBd, LNA loss 1,8dB (long cable)	
LNA NF 0,45	LNA gain 15 dB
Coax loss 1db	RX NF 1,0
Side lobes 10K	Back lobe 40K
RX BW 50Hz (good CW operator!)	
TX 500W	TX loss 1,6dB

Bandwidth correction: WSJT measures in bins of 0,66Hz. So normally the used bandwidth in WSJT W is not 1. Typically I saw in my qso's around 2-3 what means a correction of about 6dB. See the article from Rex VK7MO in DUBUS 1/2007 which shows how to use WSJTs echoes mode to check your station. To be continued.

Interesting links

VK3UM Libration Calculator.

What would be most helpful is to confirm the predicted values. All you need is equipment capable of displaying your echo (or others) through a waterfall display and save such for analysis. We would appreciate your participation, please. The higher the frequency the better, but those who operate 70cm and higher will be able to 'see / hear' the effect. Additionally, **any time you may be able to take a measurement will be helpful** to check predicted against calculated values. Pse forward your findings to G3WDG charlie@sucklingfamily.free-online.co.uk For further info re times please refer to my LibCalc program on www.vk3um.com 73 Doug VK3UM

Libration rates have been included in the program WSJT for some time now. They are also in the MoonSked program by David Anderson, GM4JJJ. For theory look at power point sheet www.physics.princeton.edu/pulsar/K1JT/refs.html and click on references 14 and 15. Libration-induced Doppler spreads for self echoes and for the station you're working can be read directly from the "Astronomical Data" screen of **WSJT 9.0**. The values labeled "Spread" are the maximum limb-to-limb broadening, in Hz. Those labeled "w50" are the width (in Hz) that will contain about 50% of the echo power.

Who maybe in concern about **high pwr MW combiners**, can find on www.ok1kir.cz constructional details of pwr combiner (3dB hybrid) used in our new 1kW 23cm SSPA. It is a modified Kathrein professional design (moved up in frequency even up to 13 & 9cm bands) with excellent wide-band performance.

Andreas DJ3JJ updated his **RX performance** web side. Look at: www.do9bc.de/index.php?id=28 . There is a completely new layout of the main page now.

The latest **VK3UM EME Calculator** update (Ver 7.08) is now available. Download direct from www.vk3um.com Regards Doug VK3UM

At this link you can find the Results of Italian EME Contest and more. 73 GL de Enrico I5WBE www.eme2008.org/ari-eme/contest.html

Silent Key

Sad news but unfortunately predictable. On 01/01/11 at 01:01 Claude F3VS became a SK. I lose a friend from my youth who had transmitted the virus of ham radio and EME. 73 Philippe F2TU

Final

We just starting again a discussion how to get more CW activity on EME, especially on 432. See CT1DMKs very nice statistic on the ARRL EME contests over the years on www.qsl.net/ct1dmk/eme_stat.html . As we can see the CW activity had its best 1993 on 432, then the numbers get smaller every year. This was way before we had digital modes. One of the reasons is the growing numbers on 23cm and up, they are mostly moving from 432 up to higher bands. Other reasons are maybe other possibilities / hobbies, more workload in qrl, growing QRM level, getting older and problems to put up a setup. Today its much much harder to run EME in a big city with a rented apartment including an antenna permission. But the growing number of new licenced HAMs in the US shows that there is potential. If we can win the younger ones we can hope for more activity. Why is CW activity so low? I believe the main reason is that most of the EMERs are hunting #, DXCC or squares. Many of the active EMEers are in some toplists or participate a contest. So we all try to look for something new and the big number of new stations in the last years was only in digital. And we all know calling CQ for hours with no reply is not funny. Maybe it needs a critical number of active stations to keep activity alive. Regarding 432 random activity we passed that number more then 10 years ago. So what to do? Join the activity periods and attend in the contests to show that 432 CW is not dead.

73 Bernd DL7APV

Operating V/U/SHF EME legally from abroad, by Michael Kohla, DL1YMK

With the increasing interest in EME-DXpeditioning, to which the author of this lines is proud to have contributed significantly in recent years, it seems to be appropriate to spend a few words on the legal prerequisites that have to be met when operating EME from foreign countries.

You think this is all self-understanding and clear to everyone? Well, this is what I thought myself for quite some time, but having talked to people planning or setting out for EME-DXpedition-trips lately have indicated some, let's say it polite, disorientation regarding this issue.

Especially in context with the CEPT-regulation there is a lot of adventurous misbelieving, especially amongst European operators.

The CEPT-regulation T/R 61-01 with its revision in 2003 (removing morse code requirements) in its core statement just enables ham radio operators as residents of a CEPT-participating country to operate from another CEPT-territory for a limited period of time without having the need of obtaining an individual temporary license from the telecom authorities in the country visited.

So far, that's very convenient - but the regulation also quotes verbally in appendix 1, sent. 2.2: 'The license holder shall observe the provisions of the ITU Regulations, this Recommendation and the regulations in force in the country visited. Furthermore, any restrictions concerning national and local conditions of a technical nature or regarding the public authorities must be respected...'

That simply means, when going on an EME-DXpedition within CEPT-countries, you have to comply with the national power limits and frequency allocations - simple as that - and of course, you may not 'import' your own national regulations, as surprisingly many people seem to think. In other words: with the exception of only very few countries with liberal power limits in the V/U/SHF-range (unfortunately mostly not preferred EME-destinations), the EME-DXpeditioner must apply for a special individual high power license as an inevitable basis for a legal operation, which at least in my eyes, is a precondition for honestly bringing up your DXCC-standings. The original wording of the CEPT-regulation can be found at:

www.erotodocdb.dk/doks/implement_doc_adm.aspx?docid=1802

Outside the CEPT-countries the EME-DXpeditioner without any exception has to apply for a temporary high power permit, which - who would know it better than the author - in most cases is very hard to get, and even sometimes impossible. For example the DL1YMK/CX operation from Uruguay in 2008 was of course fully legal, also on 2320 MHz, as the national telecom authorities after lengthy negotiations had

agreed to a temporary permit with EME-power on a NIB-basis, providing a report about the active and passive interference situation written by the operator. This was the first time ever activation of South America on 13 cm, providing a lot of WAC's. For the very reason of power restrictions, the author had to finally drop a DXpedition to CE in spring 2010, as even after intensive efforts with the kind support of the president of the national amateur radio club as well as the Secretary General of IARU Reg. 2, I was not able to obtain an EME-suitable permit, covering the V/U/SHF-bands. The license obtained only covered - according to national regulations - HF with high power. Working with the allowed 50 W on 70 cm and only 10 W on 23 cm even for JT seemed definitely useless to me. According to my own experience, applying for a dedicated EME-permit, especially comprising frequencies above 144 MHz, in many cases is not successful, although consuming a hell lot of time, thus removing a whole number of most wanted destinations from the list - sad, but true! In this context I recently learnt from a good friend of mine, who for more than a year tried to get a license in a very interesting European country, finally cancelled the planned EME-DXpedition, because the local authorities refused to grant the necessary EME permit.

On the other hand, from my point of view at least, it is a matter of ham spirit and personal integrity to perform EME-DXpeditions only from rare destinations, if a legal basis really is given. If not, it would be extremely unfair for an unlis DXpedition to work EMEer's all around the globe, leaving them in good believe that their accomplished contacts were in compliance with national regulations and thus bring up their worked country score. Another consequence of an illegal operation may be the refusal of EME-permits for future DXpeditions by the local authorities.

Hence, it is clearly the responsibility of the DXpeditioner to provide a sound basis for the enterprise, thus also not alleviating the efforts of those, who are honest with themselves and their fellow EMEer's. Nevertheless, I see a strong tendency that those stations eagerly scoring DXCC's have only little interest in the facts mentioned above these days - just the numbers count. Well, everyone has to handle this for himself.

The IOTA-program on HF, to which I contributed many years ago with DXpeditions to rare Aleutian and Mexican islands, has found a very strict, but reasonable way to make sure that only valid contacts are made with those rare islands: the DXpedition station has to minutely prove the legal compliance by disclosing all licensing documentation details to a designated expert, in order to provide a comparable basis for everybody involved in the program. Hopefully, this is not necessary with EME? Just thinking... 73 Michael

WORLDWIDE EME CONTEST 2011

sponsored by DUBUS and REF

CW / SSB only

dedicated to F3VS, Claude Descat (SK)

The WW EME contest is intended to encourage world-wide activity on moonbounce. Each different call prefix forms a multiplier.

1. Contest Dates & Bands

First weekend	March 12 / 13	00 - - 24 UTC	144 MHz & 10 GHz + up - CW/SSB
Second weekend	April 9 / 10	00 - - 24 UTC	432 & 3.4 GHz - CW/SSB
Third weekend	April 30 / May 1	00 - - 24 UTC	5.7 GHz - CW/SSB
Fourth weekend	May 7 / 8	00 - - 24 UTC	2.3 GHz - CW/SSB
Fifth weekend	June 4 / 5	00 - - 24 UTC	1.2 GHz - CW/SSB

2. Sections and Awards

QRP 144MHz <100kW EIRP 432MHz <400kW EIRP 1296MHz <600kW EIRP
but no separate QRP/QRO categories

QRO On 144, 432 and 1296MHz, stations with EIRP equal to or greater than stated above.

CW/SSB All QSOs in CW and/or SSB mode – no other modes used

MULTI Multi-OP is >1 OP – but no separate category

Multi-operator and QRO stations will be highlighted in the general classifications. All QRP/QRO band winners and QRP/QRO multiband winners will receive a year's free subscription to DUBUS magazine. The multiband section contains weekends 1, 2, 3, 4 and 5. In each band/section certificates will be sent to the first 10 places.

3. Rules

3.1 For the purpose of the contest only one scoring per valid QSO with the same station can be logged in each band.

3.2 During the European EME Contest dates & times, communication via the Earth-Moon-Earth path is the only type of communication permitted by participants and stations worked.

3.3 **During Contest time**, it is not allowed to use other communications medium such as internet or packet radio, to self spot, announce CQ frequency, make skeds, exchange any QSO progress info, confirm whether the QSO was valid or not.

3.4 Spotting other stations for activity reasons, is permitted.

3.5 If stations participating in the Contest choose to use 'real time' communication outside the Earth-Moon-Earth path so as to solicit contacts for themselves, they are permitted to do so, only outside their Contest participation time. In such a case, re-entering the Contest for the remaining of their respective weekend, moon time, is not allowed.

3.6 Stations participating in the Microwave bands (2.3GHz and above) are permitted to announce their time plan of proposed band segment activity, during times when they have no moon visibility.

3.7 Stations deviating from the rules are not eligible to submit logs for the European EME Contest.

4. Contest Exchange

For a valid EME QSO, **both** stations must have copied all of the following:

- 4.1 Both callsigns from the other station
- 4.2 Signal report from the other station (using TMO procedure or RST)
- 4.3 R, from the other station, to acknowledge complete copy of 4.1 & 4.2

5. Logs

Logs must be separate for each band, and should be in normal "logbook" format.

Top line: Your callsign, Band

Each QSO: Date/Time, Callsign, Report sent, Report received, Points, Multiplier

Bottom line: Total points, Total multipliers, Total claimed score.

6. QSO Points

100 points for each random QSO completed on 144MHz – 10GHz.

10 points for each sked QSO completed on 144MHz – 10GHz.

7. Multipliers

Each different call prefix is a multiplier (e.g. DL1, DK9, SM2, S51, S54, G6, KM5, W5, JA6, VK4, WA6, K6, PA1, PE1, etc). See example of WPX Contest rules for further details on prefix multipliers.

8. Total Scores

Single band score = [Total of QSO points] * [Total of multipliers].

There will be one QRP winner and one QRO winner on each band.

Multiband score = [Total sum of points on 144-10GHz] * [Total sum of multipliers on all bands]

Multiband stations will also be listed as an entry on each separate band worked, and can also win single band awards.

9. Contest Entries

Copy of the log for each band with details of points, multipliers and total points.

The following information **MUST** also be included for each band:

1. Output power, transmit cable loss, antenna type and gain
2. Categories: QRO/QRP - single/multi operator
3. Start time and end time of Contest participation time
4. Name(s) of all operators
5. Grid locator.

Other info is welcome: Comments, conditions, station details, photographs, etc.

10. Sending Your Entry

Contest entries **MUST** be sent no later than 20 days after the end of the last contest weekend (i.e. in the mail or e-mail by 25 JUNE 2011).

Mail address: Joachim Kraft, Gruetzmuehlenweg 23, D-22339 Hamburg, Germany

You can e-mail your contest entry in any common format to: funk-telegramm@t-online.de

All email entries will be acknowledged within one week. If you have not got any acknowledgement, please send your entry again.

For further questions contact: funk-telegramm@t-online.de

Good Luck in the Contest!

For REF: Patrick Magnin, F6HYE

For DUBUS: Joachim Kraft, DL8HCZ/CT1HZE

2m EME News

Upcoming expeditions and activities

ZA Albania on 2m EME

OK1DFC etc plan a big EME and MS expedition on 2m, 70cm, 23cm and up in May. Call may be ZA1EME.

GD Isle of Man

MD/DH7FB & DF2ZC will be qrv from April 10 to 13 from IO74RH on 2m EME in JT65. Rig 2 x 2M9 and GS35. Working frequency is 144,114 MHz and 'MD/' is always tx 1st. Updates on <http://twitter.com/dpexpeditions>

Kasakhstan by UN/DL9LBB

Hans has moved to Astana in Kasakhstan in autumn 2010. He got his licence and plans to put up antennas for 2m and 70cm EME in spring. His grid is MO51 and he will stay there for about 3 years.

New DXCC Bonaire on 2m EME for 1st time

From April 4-11. PJ4NX, PJ4/PA3FPQ, PE1LWT + PE1L will be qrv on 2m EME in JT65 with 4 x 9 Ele and 1 KW. QRG 144.124 MHz PJ4 first. Loc. FK52.

DUBUS 2m CW EME Activity Contest

April 2 0430-0730 GMT & 1300-1600 GMT

May 1 0430-0730 GMT & 1230-1530 GMT

June 26 0100-0400 GMT & 0900-1200 GMT

July 23 2300-0200 & July 24 0800-1100 GMT

Aug 27 0330-0630 & Aug 28 1300-1600 GMT

Dec 3 1200-1500 GMT & 2030-2330 GMT

QRG: 144.040 – 144.060 MHz. Pse check for more info, updates and results on the web at: <http://www.sm2cew.com/dubus-aw.html>

DUBUS 2m CW EME Activity Contest – Reports

SV10AA (Costas): I was QRV and really enjoyed working both windows of the December Dubus 2m ATP. Conditions were not good here on SV-Land on the start but improved as time passed. I had complete QSOs with SV1BTR RO/O, YO2AMU RO/O and KB8RQ RO/O. Also copied DK3NG and VE1KG but no QSO was established. Outside ATP times, on sked, I worked IK2DDR. 73 from Costas, SV10AA

SV10AA (Costas): I was QRV during both legs of January 2m CW EME ATP and very much enjoyed my participation. During the first leg, I found good activity and conditions here to be better than usual. I could hear several stations calling CQ but due my usual high QRM and QRN I could not get all their callsigns. At the end of the first leg and after 4 successful QSOs with OZ1HNE RO/O, OK1MS RO/O, SV1BTR RO/O and YO2AMU RO/O, I was very optimistic. But during the second leg conditions here got worse with especially high QRN, so no further QSOs were established unfortunately. Heard but no QSO was established: DK3EE, W3TWX and DK3NG. 73 from Costas, SV10AA

REF / DUBUS EME Championship 2010 - Results by DL8HCZ

144 MHz Digital only

Place	Call	Points	QSO	Multi	Setup
1	RU1AA	17316	148	117	16x14, GU93
2	K5QE	15732	138	114	8x18, 1K5
3	RK3FG	15707	139	113	4x15, GS35
4	IK1UWL	11737	121	97	4x14, QRO
5	UA3PTW	10374	114	91	8x15, 1K2
6	DM3X	7968	96	83	3x10, GS35
7	I2FAK	6942	89	78	16x19, 1K2
8	JR3REX	4544	71	64	4x17, 2kw
9	OM3BC	3960	66	60	4xXP32, KW
10	WZ5Q	3944	68	58	2x17, 1kw
11	EM5U	3472	62	56	2x2.5l, 500w
12	K14TZ	2650	53	50	2x20x, 8877
13	SP2NJI	2300	50	46	4x8, 500w
14	F8KFX	2254	49	46	2x12el, 400w
15	EA6VQ	1804	44	41	8xly
16	JE1TNL	1763	43	41	2x28x, QRO
16	UN9L	1763	43	41	2x16, 650w
18	VK2KU	1476	41	36	4x12, 1kw
19	UY9VY	1332	37	36	2x16, 1kw
20	W6YX	1224	36	34	1x5l, 350w
21	YU7XL	1188	36	33	16x8, 700w
22	EB1DNK	1054	34	31	4x4.5l, GS35
23	DF0SY	990	33	30	2x8, 750w
24	K5DOG	961	31	31	2x9, kw
25	CT1HZE	840	30	28	4x11el, 2kw
26	DH1WM	756	28	27	2x17, 500w
26	F9HS	756	28	27	1x3l, 1kw
28	I3EVK	702	27	26	4x19, 700w
29	DL2FCN	676	26	26	2x11, 750w
30	DL7UDA	676	26	26	
31	G4ZTR	598	26	23	4x8el, 400w
32	YO3DMU	441	21	21	2x16, 1kw
33	AE3T	225	15	15	12el, 500w
34	G4KLA	196	14	14	1x11, 1kw
35	ZS6OB	156	13	12	4x16el, 100w
36	UA9JMB	100	10	10	
37	AK4T	25	5	5	2x8el, 200w

Checklog: DL4KUG

Congratulations to Alex, RU1AA and Serge, RX1AS for winning the 2010 Championship! Top scores are way up compared to 2009. Also there were about 50% more participants in 2010. Looks like that this concept is quite OK.

EUROPEAN EME Championship 2011 – 2m

The Digital 2m EME Championship will take place on July 30 and 31, 2011 from 00-24UTC, both days. We hope for good participation!

Also "small" logs are welcome.

Pse send entries and logs to funk-telegramm@t-online.de

Transequatorial Propagation on 6m and 2m over South America

by FM8DY, Jean-Louis (Grid: FK94MO, Martinique Island)

DD/MM/YY	6m	QSO (Q) or reception (R) on 6m	Solar Flux	2m	QSO or reception on 2m (!)
30.08.10	Q	PP5, PY5	75	-	
04.09.10	Q	PY5	82	-	
11-12/09/10	Q	PY2, PP5, PY5	78	-	
17.09.10	Q	PY5, PU5, PY4	82	R	
18.09.10	Q	PP5	82	R	PP5XX 419
19.09.10	Q	PY2	81	Q	PP5, PY5 via repeater, PU2YEN direct GG66MV
22.09.10	Q	PY2	85	R	Repeater
23.09.10	R	PY2	84	-	
27.09.10	Q	PY2	83	-	
29.09.10	Q	PY2, PP5	91	-	
30/09-11/10	Not	QRV			
12-13/10/10	Q	PP5, PY2, PY4, PY5, PY3, CX1, LU4, ZP6	75	Q	PP5, PY5
13-14/10/10	Q	LW5, LU1	78	Q	PP5, PY2 GG48, PY5
14-15/10/10	Q	PY0FF/B, ZD8VHF/B	80	Q	PP5
15-16/10/10	Q	PY2, PS8, PU2	82	R	PY repeater
16-17/10/10	Q	PY5, ZD8VHF/B, PS8RF/B	87	Q	PP5, PY5
17-18/10/10	Q	LU1, PY5, PU2, PS8	84	Q	PY5, PU2, PP5
18-19/10/10	R	PY0FF/B, ZD8VHF/B	91	-	
19-20/10/10	Q	PY0FF (QSO), ZD8VHF/B, LU3, PY2, PY5, PU5	87	Q	PP5, PY5 XX
20-21/10/10	Q	PY2	84	Q	PY5, PU2, PP5, PU5
21-22/10/10	Q	LU3, LU7, LU9	84	Q	PY5, PP5, PU2 XX
22-23/10/10	Q	PY2	82	Q	PY5, PP5, PY2
23-24/10/10	Q	PY2, PY5, PU7, PU2, PP5	84	R	PY2
24-25/10/10	Q	CA3, LU3, LU1	82	Q	PP5, PY5, PU5 XX
25-26/10/10	Q	PY5, ZD8VHF/B, OA4B/B, CA3, CE3	86	Q	PY5, PP5, PU5, PU2
27-28/10/10	R	PP2SIX/B, PY2XW/B, LU7FTF/B	88	Q	PY5 XX
29-30/10/10	R	LU7FTF/B	86	R	PY repeater XX
Antennas		Preventively down for			hurricane Tomas
05-06/11/10	-	Most time on 2m the following days	83	Q	PY5, PU2
06-07/11/10	-		89	Q	PP5, PY5, PY2, UT1FG/MM GG64
07-08/11/10	-		85	Q	PP5, PY5, PY2, UT1FG/MM GG63, PY2 GG48
08-09/11/10	-		84	Q	PY5, PP5, UT1FG/MM GG64 XX
09-10/11/10	-		84	Q	PY5, PY2, UT1FG/MM GG63 XX
10-11/11/10	-		86	Q	PY5, PY2, UT1FG/MM GG64 XX
11-12/11/10	-		85	Q	PY5 XX
12-13/11/10	-		85	R	PY2
13-14/11/10	-		85	Q	PY5 XX
14-15/11/10	-		86	Q	PP, PY2, PU2, UT1FG/MM GG64
15-16/11/10	-		91	Q	PY5
16-17/11/10	-		92	Q	PY5, PP5, UT1FG/MM XX
17-18/11/10	-		91	Q	PY2, PY5, PU2, PP5
18-19/11/10	-		87	Q	PY5, PU2, PP5
19-20/11/10	-		84	Q	PY5
20-21/11/10	-		80	Q	PU2, PY5, UT1FG/MM GG54
21-22/11/10			78	Q	PY2, PY5, PP5, UT1FG/MM GG54
22-23/11/10			75	Q	PP5, PY3IOD GF49 5077km ODX, PY5, PU2 XX
23-24/11/10			75	Q	PY5
24-25/11/10			76	R	PY5 repeater
25-26/11/10			78	Q	PY5, UT1FG/MM GG62, PP5 XX
26-27/11/10			76	Q	PY5LF
27-28/11/10			77	R	PY repeater
28-29/11/10	R	CX1CCC/B, CX1AA/B, LU7FTF/B	80	R	PY repeater
29-30/11/10	-		83	Q	PY5 strong
30-01/11-12/	Q	OA4TT, LU6, CE3, PY4	86	R	PY repeater
01-02/12/10	Q	PY2	87	R	PY repeater
02-03/12/10	Q	PP5	87	Q	PP5
03-04/12/10	Not	QRV	87		
04-05/12/10	Q	PY2, CW3, LW3, ZD8VHF/B, CX1CCC/B	87	R	PY repeater

05-06/12/10	Q	CW3,PY2	88	Q	PY5
06-07/12/10	Q	LU8DIO, PY2	89	Q	PP5
07-08/12/10	-	55,250 TV only, weak	87	-	
08-09/12/10	R	LU1IBL/B, LU7FTF/B, CE3AA/B, CX1AA/B	87	R	PY5GRU repeater
09-10/12/10	-		87	R	PY5GRU repeater
10-11/12/10	R	PY4, LU8, LU4	88	R	PY5GRU repeater
11-12/12/10	Not	QRV	87		
12-13/12/10	-		89	R	PY5GRU repeater weak
13-14/12/10	-		88	R	PY5GRU repeater
14-15/12/10	-		90	Q	PY5
15-16/12/10	Q	PY2, CE1, LU5	87	Q	PY5, PU2
16-17/12/10	Q	CE4	84	R	PY5GRU repeater
17-18/12/10	Q	PY2, hrd AC4TO Es	82	R	PY repeater
18-19/12/10	Not	QRV			
19-20/12/10			81	R	PY repeater
20-21/12/10			78	Q	PP5
21-22/12/10			78	R	PY5GRU repeater weak
22-23/12/10			78	R	PY5GRU repeater
24/12-31/12	Not	QRV			
01-02/01/11	Q	LU2, LU4	91	-	
02-03/01/11	Q	PY5, PU2, PY2, CX1CCC/B, CX1AA/B, LU5EGY/B	91	R	PY5GRU repeater weak
03-04/01/11	R	TV 55,250	92	R	PY5GRU repeater weak
05-06/01/11	Not	QRV			
07-08/01/11	Q	PY2, PU3SIX/B, CX1CCC/B, CX1AA/B, LU5EGY/B	86	R	Many PY repeaters strong
08-09/01/11	Q	PY1, PU3SIX/B	85	R	PY5GRU repeater
09-10/01/11	R	PU3SIX/B, LU1IBL/B	83	R	PY5GRU repeater
10-11/01/11	R	PU3SIX/B, LU1IBL/B	83	Q	PY2EQJ Via Repeater 145,370
11-12/01/11	R	PU3SIX/B	83	Q	PY5
12-13/01/11	R	TV 55,250 weak	80	-	
13-14/01/11	R	LU1IBL/B	80	R	PY repeater weak
14-15/01/11	R	TV 55,250	79	R	PY FM simplex 146,510
15-16/01/11	Q	PY2, PU2, CE1B/B, CE3AA/B	80	Q	PU5
16-17/01/11	Q	PY2, CE3AA/B, LU7FTF/B	80	-	
17-18/01/11	R	OA4TT/B, OA4B/B	82	-	
18-19/01/11	R	LU1IBL/B, LU2MCA/B, LU7FTF/B	81	Q	PY5LF strong, PU5, PY5
19-20/01/11	Q	PY2, LU1IBL/B, PU3SIX/B, CX1CCC/B, LU8DZE/B	82	R	PY5GRU repeater weak
20-21/01/11	R	TV 55,250	81	R	PY5GRU repeater weak
21-22/01/10	Not	QRV			
22-23/01/11	R	LU1IBL/B, LU7FTF/B	88	R	PY5GRU repeater weak
23-24/01/11	R	LU1IBL/B, LU7FTF/B	84	R	PY5GRU repeater weak
24-25/01/11	R	LU1IBL/B, LU7FTF/B, LU5EGY/B	83	R	PY5GRU repeater good
25-26/01/11	R	LU1IBL/B	81	R	PY5GRU repeater weak
26-27/01/11	R	LU1IBL/B	80	Q	PU5RAP in FM simplex
27-28/01/11	R	LU1IBL/B, CE3AA/B, OA4TT/B, LU7FTF/B	81	-	
28-29/01/11	-		81	R	PY repeaters+simplex FM strong
30-31/01/11			83	Q	PY5, PU5, PU2 XX

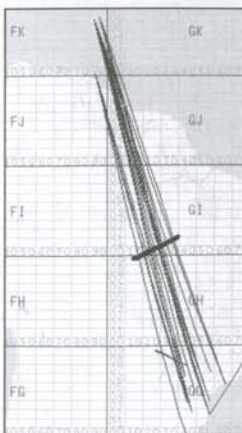


Table: TEP observations by FM8DY, FK94MO, in autumn and winter 2010/2011. 2m rig: 100w & 6 el. Yagi. One can see clearly that a "better" flux of >90 does not necessarily means good TEP on 2m and a "poor" flux of <80 not necessarily poor TEP on 2m. FM8DY had 81 days (!) with 2m TEP between 18. Sept 2010 and 31. January 2011 (136 days – 30 days when he was not QRV = 106 days qrv). That is 76% with 2m TEP of the days when he was QRV!

XX = 2m TEP QSOs reported between YV4 area (FK60) and LU (GF05) area added by CT1HZE

Map: 2m TEP QSOs on Nov 20th 2010 (map by ON4KST, tnx) with typical paths in the 4500 km range from Martinique (FM), Guadeloupe (FG) and Trinidad (FK90) to Brazil (PY3, PY2, PY5), the black line in the centre of the path is the geomagnetic equator

Tropo Reports

2m, 70cm and up

Editor: Michael M. Dienel, DG7SFL
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2m

DL3YEE wkd on 144 MHz

Date: 12 Oct 2010

Time	Call	Loc	QRB	Mode/Remark
07:46	GW8ASD	IO83LB	786km	

UR5LX in KO70WK wkd on 144 MHz

Date: 07 Dec 2010

Time	Call	Loc	QRB	Mode/Remark
16:46	UN7AX	LN53NP	1379km	

70cm

DL3YEE wkd on 432 MHz

Date: 12 Oct 2010

Time	Call	Loc	QRB	Mode/Remark
07:42	GW8ASD	IO83LB	786km	
20:22	G8XVJ/P	IO93AD	715km	
20:26	E18IP	IO63NR	1044km	
20:29	G0AJJ	JO02QT	490km	
20:33	G7HOA/P	IO93AD	715km	
20:35	G0EHV/P	IO84XT	757km	
20:40	GM4BYF	IO85JV	869km	
20:48	G4KUX	IO94BP	741km	
21:00	OK1VVT	JO60VR	398km	

G3XDY wkd on 432 MHz > 600km

Date	Call	Loc
14 DEC 10	DF9IC	JN48
11 JAN 11	DL6NAA	JO50
11 JAN 11	SK7MW	JO65
11 JAN 11	DL0VV	JO64
31 JAN 11	DL5DWF	JO71
01 FEB 11	DJ7GK	JN68

23cm

DL3YEE wkd on 432 MHz

Date: 12 Oct 2010

Time	Call	Loc	QRB
07:57	GW8ASD	IO83LB	786km

G3XDY wkd on 1296 MHz > 600km

Date	Call	Loc	
16 NOV 10	DF9IC	JN48	
16 NOV 10	DL0VV	JO64	
16 NOV 10	OZ1FF	JO45	
16 NOV 10	OZ9KY	JO45	
11 DEC 10	F6FHP	IN94	
21 DEC 10	DF9IC	JN48	
18 JAN 11	DL0VV	JO64	
18 JAN 11	OZ1FF	JO45	
18 JAN 11	DF9IC	JN48	
31 JAN 11	DL5DWF	JO71	New square
31 JAN 11	OE5VRL/5	JN78	1012km

OE5VRL/5 in JN78DK 866m asl wkd >800 km

Time	Call	Loc	QRB	Mode/Remark
Date: 31 Jan 2011				
22:33	G3XDY	JO02OB	1012km	SSB
Date: 1 Feb 2011				
15:12	F6DKW	JN18CS	888km	SSB
15:39	F1RJ	JN18AT	900km	SSB
18:13	F5DQK	JN18GR	864km	SSB
18:50	F5HRY	JN18EQ	876km	CW

OZ1FF in JO45BO wkd on 1296 MHz

with QRB > 600 km

Date: 16 Nov 2010

Time	Call	Loc	QRB	Mode/Remark
18:45	DF9IC	JN48IW	742km	
18:50	SM0ERR	JO89WJ	718km	
19:47	GM4CXM	IO75TW	781km	
20:06	G3XDY	JO02OB	600km	
20:48	SK0CT	JO99BM	737km	

Date: 21 Dec 2010

Time	Call	Loc	QRB	Mode/Remark
17:50	DF9IC	JN48IW	742km	
18:00	SM0ERR	JO89WJ	718km	

Date: 18 Jan 2011

Time	Call	Loc	QRB	Mode/Remark
18:40	SM0ERR	JO89WJ	718km	
19:30	DF9IC	JN48IW	742km	
20:52	GM4CXM	IO75TW	781km	
21:04	G3XDY	JO02OB	600km	
21:31	SK0CT	JO99BM	737km	

RIG: 1,5 m dish, 145 W

13cm

F6DWG/P JN19AJ wkd:

Date	Call	Loc
01 FEB 11	DJ9PC	JN59PO 668km CW

G3XDY wkd on 2320 MHz > 600km

Date	Call	Loc
16 NOV 10	DF9IC	JN48
11 DEC 10	F6FHP	IN94
31 JAN 11	OE5VRL/5	JN78 1012km

OE5VRL/5 in JN78DK 866m asl wkd >800 km

Time	Call	Loc	QRB	Mode/Remark
Date: 31 Jan 2011				
22:33	G3XDY	JO02OB	1012km	SSB

OZ1FF in JO45BO wkd on 2320 MHz

with QRB > 500 km

Date: 28 Dec 2010

Time	Call	Loc	QRB	Mode/Remark
17:03	SM7LCB	JO86GH	529km	
18:43	DF9IC	JN48IW	742km	
20:19	SA4Z	JO79OF	584km	
20:39	SM0ERR	JO89WJ	718km	
Date: 18 Jan 2011				
19:50	DF9IC	JN48IW	742km	

RIG: 1,5 m dish, 200 W

6cm

G3XDY wkd on 5760 MHz > 300km

Date	Time	Call	Loc
31 JAN 11	22:38	OE5VRL/5	JN78 1012km

OZ1FF in JO45BO wkd on 5760 MHz**with QRB > 400 km**

Date: 28 Dec 2010

Time	Call	Loc	QRB	Mode/Remark
17:38	SM7GEP	JO77IP	462km	

Date: 25 Jan 2011

Time	Call	Loc	QRB	Mode/Remark
17:38	SM7GEP	JO77IP	462km	

RIG: 1,5 m dish, 7 W

3cm + up**F2CT/P IN93HG wkd on 24048 MHz:**

5 FEB 11 EA2BCJ/P IN93GF 59/59

G3XDY wkd on 10368 MHz > 300km

Date Call Loc

11 DEC 10 DL5EAG JO31

OE5VRL/5 in JN78DK 866m asl wkd >800 km

Time	Call	Loc	QRB	Mode/Remark
17:29	F6DWG/P	JN19AJ	900	CW 559 559
17:55	F6DKW	JN18CS	888	CW 559 559

Date: 1 Feb 2011

17:29 F6DWG/P JN19AJ 900 CW 559 559

17:55 F6DKW JN18CS 888 CW 559 559

OZ1FF in JO45BO wkd on 10368 MHz**with QRB > 300 km**

Date: 16 Nov 2010

Time	Call	Loc	QRB	Mode/Remark
21:14	SM6ESG	JO67CC	302km	

Date: 23 Nov 2010

Time	Call	Loc	QRB	Mode/Remark
19:37	SM6AFV	JO67GQ	356km	
20:39	SM7ECM	JO65NQ	314km	

Date: 2 Dec 2010

19:37 SM6AFV JO67GQ 356km

20:39 SM7ECM JO65NQ 314km

Date: 25 Jan 2011

Time	Call	Loc	QRB	Mode/Remark
19:42	SM6AFV	JO67GQ	356km	
20:31	SM7ECM	JO65NQ	314km	
20:57	SM7GEP	JO77IP	462km	

19:42 SM6AFV JO67GQ 356km

20:31 SM7ECM JO65NQ 314km

20:57 SM7GEP JO77IP 462km

Date: 25 Jan 2011

Time	Call	Loc	QRB	Mode/Remark
21:33	SM7ECM	JO65NQ	314km	

21:33 SM7ECM JO65NQ 314km

RIG: 0,65 m dish, 4 W

OZ1FF in JO45BO wkd on 24048 MHz

Date: 16 Nov 2010

Time	Call	Loc	QRB	Mode/Remark
21:18	SM6ESG	JO67CC	302km	

21:18 SM6ESG JO67CC 302km

TNX to ALL for submitting reports!

The 23cm MS-Burst received by OK1DFC on
4.1.2011 at 10:56z from RW7A:

105600 8.0 280 2 26 -38

7BW7P GG1DFC7RW7A7OK1DFC RW7A OK1DFC RW

New 24 GHz Tropo World Record!

On February 7th 2010 in the morning hours there were good Tropo condx between southern Germany and the Paris area in France. DL7QY JN59BD and F6DKW JN18CS worked on 10 GHz with almost endstop signals around 09:30 UTC.

Then CW tests were made on 24.048 GHz and at 0956 a complete 2-way CW QSO was made where 529 reports were exchanged. The distance is a fantastic 579 km. This is a new Tropo word record on the 1.2cm band! F6DKW was running 1.5W RF into a 50cm dish and DL7QY 1 W RF and a 70cm dish.

Claus reported that the humidity dropped in the morning down to 26% RF in his QTH and a humidity map from the French WX office showed low humidity for most of the path between DL and F. See the front cover picture for the path and antennas.

The former world record was set in the U.S. on Sept 7th 2002 between W5LUA (EM13qc) and WW2R/5 (EM41hc) over 543km.

Congratulations to Maurice and Claus for this outstanding QSO and tnx to them for submitting reports and pictures.

Meteor Scatter on 23cm?!

After the fantastic Aurora QSOs and observations on 23cm (see last DUBUS issue 4/2010 on page 128), recently some amateurs also made experiments on 23cm with Meteor Scatter. Commercial radar data suggested already quite some time ago that 23cm MS should not be impossible. Also the experiences from the signal strengths and burst durations on 70cm suggested that on 23cm there should come through at least "something". In Perseids 2010 F6FHP made a 23cm MS test with an OK2 station and reported a ping that was not decoded. In Quadrantids 2011, on January 4th, OK1DFC (JN79) made a short FSK441 test using his 10m EME dish and 1.5KW with RW7A who was using 300w and a 3m dish (Loc. KN95NA) and received 3 pings and a "real" burst (280ms/2dB) that was decoded. See screenshot capture below. RW7A also received one burst (160/1) within the 15 minute test at about 11 UTC. The distance between the two stations is 1921km. Although a QSO was not completed it should be only a question of time until the first 2 way MS QSO will be completed on 23cm. Who will make history?

4m News

Reports, Expeditions, Infos

Editor: Joachim Kraft, DL8HCZ

Poland on soon 70 MHz

Years ago it has been already announced but now it looks like that it will happen: Polish radioamateur will get the range from 70.1 to 70.3 MHz on a secondary base with max. 20 W EIRP. Rumors say that the release could be as early as March. Several SP stations are ready equipped already and biting nails. For sure many first ever QSOs will happen on the first days after the band is free.

Iceland: 2 more years on 4m

TF amateurs can get special licences for the range from 70.0 to 70.2 MHz also in 2011 and 2012. The previous licences expired on Dec. 31st 2010, although it looks like that there has not been a single ham active yet on 4m in 2010.

A miracle in the Netherlands?

It looked like that the Netherlands could be the 4m loser like it has been Hungary on 6m. But no! In December surprisingly there appeared a proposal from the ministry in charge where a secondary allocation of the range 70.0 to 70.5 MHz is suggested for amateur radio use. The necessary formal procedures could be finished already this spring, so we could expect our Dutch friends soon on the band. There are already several stations ready for TX. The first ever PA to DL took place already back in 1957 when both countries had allocations, so DL's can concentrate on the SPs.

4m NA beacon in 2011

WA1ZMS is undertaking another effort to get an experimental licence again for a 4m CW beacon for summer 2011. In summer 2010 his 4m beacon WE9XFT was on air on 70.005 MHz. The only station who heard it from EU was CT1HZE the only one from AF was EA8/DL3GCS. So hopefully there will be more reports in 2011 as it is very unlikely that there will be a 4m beacon again on air from NA in 2012.

More 4m beacon news

LA5VHF/B 70.065 MHz, JO48AD, 10W, 3 el. QTF 180, is new on air since February 2011.

OE5QL/B 70.045 MHz, JN78CJ, OE5MPL got the licence renewed and will continue operation until August 31st 2011.

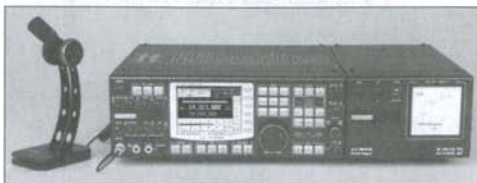
New Equipment

DB6NT offers a high class transverter with 20w output also for 4m. A separate 12V PA module (MKU PA 70 HY) with 30 to 40W is also available from DB6NT. Input power is 20 to 40mW. See DB6NT.com



There is also a rumor that a chinese manufacturer will offer a 4m transceiver soon. May be only FM, but who knows...

So far only transceiver that covers the 4m band by default in it's specifications is the Hilberling PT8000 from Germany. This top level transceiver covers all HF bands, 4m and 2m and delivers 200W output even on 4m and 2m.



PT8000 Transceiver incl. 4m & 6m band with 200W

6m News

Reports, Expeditions, Infos

Editor: Joachim Kraft, DL8HCZ

DX on 6m in December / January

A Highlight in January 2011 was a QSO on the 11th at 0108UTC between ZL1RS (RF64NS) and KC0RF (EN32CG) in CW over 12.724 MHz. There was strong Es from W9/0 to W7 and we can explain the QSO best with Es hops on both ends of the path and TEP in the middle. For SURE it was NO pure F2 QSO! In Dec and January there were a few more QSOs from W6 and W5 to ZL and VK and from ZL to OA which were enabled by the summer Es multihop propagation on the southern hemisphere.

ATTENTION!

Possible 6m beacon range shift

At the Vienna conference in February 2010 it was agreed about the proposal VI10_C5_B02 "QSY of 6m beacons to 50.400 MHz". It is now on the agenda to be ratified at the IARU-R1 Sun City conference coming August. After the Vienna conference some of our members told me that they are not happy with this proposal, and that I would have to stop this at all cost. In the forum of the UK Six Meter Group UKSMG, I found some mails with a similar content. I now could write a new proposal to the effect that the beacon band should remain as is, or one that would in effect be a new bandplan for the whole six meter band. Then we would have to discuss this point again in Sun City. But before I do this, I would like to ask you other VHF Managers what you think is the best way to manage this dilemma. Please send me a mail with your opinion (hb9ttq@hb9ttq.ch).

Stefan Streif, HB9TTQ in IARU Region 1 VHF Newsletter 55

6m Beacon News

CE1BB/B 50.019 MHz, FH41, 30w, H loop, proposal

CE6B/B 50.059 MHz, FF30, 30w, H loop, proposal

CE7B/B 50.009 MHz, FE35, 30w H loop, proposal

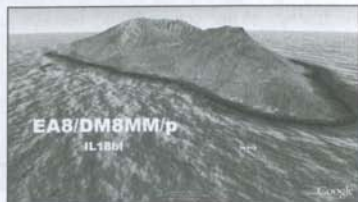
CE1B/B 50.039 MHz, FG46TL, 40w, halo, on since March 10

CE8B/B 50.049 MHz, FD46NX, on air since Feb. 2011

VA5MG/B 50.033 MHz, DO62, 30w, H-loop. On air since December 2010.

YU0ZNI 50.047 MHz, KN03WH, is the new call of ex 4N1ZNI

ZS1SIX 50.080 MHz, JF96, 10W, dipole, is on air since 12/10



EA8/DM8MM report

I was qrv from IL18BL, La Palma southwest coast on 6m with 100 W into a 5 el Konni-Yagi, free horizon from S to NW, in the end of January. Here my log:

26 Jan 2011: no DX, only local QSOs

27 Jan 2011

22:00 ZD8VHF 50,032 MHz, 519

23:30 ZD8VHF 549

23:55 ZD8VHF 539 station cl

28 Jan 2011: no DX until 23:00

29 Jan 2011

21:40 ZD8VHF 519 max 529 at 22:30

23:50 ZD8VHF 519, the whole evening with light auroral tone, modulated with approx. 100 - 200 Hz

30 Jan 2011

22:30 ZD8VHF 519

22:50 FY7THF 529 GJ35

until station closed at 23:45 ZD8VHF in the noise, FY7THF 519. Sorry, no DX QSOs made at all this time.

Will be qrv again on 6m and 2m for TEP QSOs from March 8 to 29 from same QTH. vy 73, Friedhelm, DM8MM

contests on 2m, 70cm and 23cm. Call: DR3M, loc. JO34WE.

JX Jan Mayen: An international group will be qrv as JX7VPA from July 6 to 14 also on 6m in CW/SSB. janmayen2011.org

KM27: SV1OAA will be qrv on 2m Es and EME regularly from his summer house in KM27DP in 2011.

PJ4 Bonaire: W0SD, W7XU + ON4IQ will be qrv from about mid June to July 5 on 6m with QRO. Grid: FK52.

PJ6 Saba: K5AND, W5OZI + K5TR will be qrv from 22.6-6.7. on 6m with QRO for EU. CW on 50.106, SSB on 50.118 MHz.

PJ4+PJ6 count as new separate DXCC's since Oct. 10th 2010

PJ7 St. Maarten: W6JKV + K6MYC will be qrv from June 20 to July 7, 2011 on 2m and 6m, incl. EME with QRO. Loc.: FK87. PJ7 is a new DXCC since Oct. 10th, 2010.

T32 Christmas Isl: JA1RJU will take part as an 6m EME OP in the first 2 weeks of the T32C expedition in Sept./Oct. 2011.

V2 Antigua: W9DR will be qrv from 14.-28.6. as V25DR in SSB/CW on 6m (50.115) qrv. QSL via HC.

XX9 Macau: BD5RV still plans to be qrv from Macau on 2m EME in JT65 sometime. Loc.: OL62SD.

Beacons

DK2MN/B DK2MN reports that he plans to bring his beacons back on air from 9cm & up from a new QTH in Bentheim (JO32OH) in spring.

LA7SHF 1296.945 MHz, JP20QJ, 565m a.s.l., 18w, yagi, S

OZ5SHF, 10368.900 & 24048.900 MHz, GPS locked, 25 & 7 WERP, is on air again from JO45OU since February and will change to JO45WX any day.

SV2JL/B 144.450 MHz, KN10NO, 15w, HB9CV, QTF 270. 1023m a.s.l. New on air since January 14th 2011.

S55ZMS/B 10368.105 MHz, JN86CR, new in test operation on test QRG since January 2011.

UR5LX/B 1296.990 MHz, KO70WK, 0.2w, will upgrade to 5w soon, on air since January 2011.

EA3UHF 432.481 MHz, JN01WU, 20w, loop. On air since 25. February 2010. Same QTH as ED3YAR on 1296.936 MHz. Reports pse to ea3bb@ea3bb.net

News & Comments

Editor: Joachim Kraft, DL8HCZ
info@DUBUS.de

Expeditions and DX

/MM UT1FG/MM is back from EU to the Gulf of Mexico and Mississippi area and qrv on 6m, 2m and SAT.

3D2C Conway Reef: YT1AD etc. will be qrv from Oct. 1 to 10 2011 also on 6m and 2m EME in JT65.

CN Morocco: F2CT will be qrv again from IM52 from about June 15/20 to July 15/20 2011 on 144/432/1296MHz.

DX0 Spratley, Pagasa Isl.: The announced expedition was shifted from January to may be April or later this year. Initially GM0RLZ planned to operate as DX0DX on 2m and 6m EME in JT65, but it is not sure if he will join the team again. Check updates on www.dx0dx.net

EA8, IL18, La Palma: EA8/DM8MM will be qrv from 8.-29.3. on 6m and 2m for TEP tests.

FJ St. Barthelemy: FJ/OS1T will be qrv from 14.-22.5. on 6m in SSB with 1 KW + Hexbeam. QRG: 50.115 MHz. QSL: HC.

HS Thailand: DL2LAH is qrv again until about 14.3. as als HS0ZIL on 2m via EME in JT65. Loc. OK16BQ.

JN12 EA3/DH4FAJ will be qrv from 6.-20.8. on 2m + 6m via Tropo + MS. Loc: JN12NG

JO34 DG7TG plans to be qrv again from Helgoland Isl. on May 7./8., July 2./3., Sept. 3./4. and Oct. 1./2. 2011 in the



EA3UHF and ED3YAR antennas

News

Spain on the Microwaves

Finally! Until the end of 2011 all spanish radio amateurs are permitted to use the ranges 5760 to 5762 MHz and 10368 to 10370 MHz without requesting a special permit. For the range 2320 to 2322 MHz a special licence is still mandatory, but will be granted quickly after requested. This is a major step forward as in the past only few stations in EA got special licences and nobody knew why or why not....

PLC / BPL

Latest and future PLC / BPL devices will not just cover the HF range up to 30 MHz, but also 50 MHz, 70 MHz and 144 MHz. The devices are allowed to work up to 300 MHz now. Even a 30dB notch that is requested for ham bands will not really help - if implemented at all - if you have a S9+ interference in your neighbourhood and want to work weak signals. Do you understand now why many AR associations don't encourage weak signal work any more.... but promote certain "modern" modes where you don't really hear all that noise. Thus for weak signal work it will be even more important in the future to use modern antennas which have an extremely clear pattern and low noise pick up, like the G0KSC antennas.

MMMonVHF Trophy:

"Best VHF DX Expedition" of the Year 2010

During the online voting period from 1st to 20th of January 2011 we counted more than 100 voters from 32 different countries which clicked more than 300 times for 52 different Expeditions, which were published on the Review Page of www.mmmmonvhf.de during last year. Here the results:

Category EME:

1st Place and „BEST VHF - EME Expedition of 2010“:

Rene, PE1L and Eltje, PA3CEE and Johan, PE9DX (as 3B8EME & PI4KST)

2nd place: Lance, W7GJ & Bob, ZL1RS (as 3D2LR & 3D2RS)

3rd place: DK2ZF (as CE0Y, CE2/DK2ZF & OH0/DK2ZF)

Category Meteorscatter:

2 times the 1st Place and so „BEST VHF MS Expedition's

of 2010“: Teams of US8ZAL/p & EM5U - UX5UL/p - UT6UG/p - UY5UG/p

3rd place Hermann (as FY/DL2NUD & SM/OH/DL2NUD)

The complete 2010 result and lots of comments and details are at: www.mmmmonvhf.de/trophy2010.php

DF1EQ SK

Reinhard Conrads, DF1EQ, passed away after a serious disease at age of 59 on 5th December 2010. He was a pioneer on the microwave bands, an ambitious contest operator and owner of the shf beacon DB0JL. His friends - and his wife too- lovingly called him Conny. Conny, we will miss you! Our deepest sympathy is with his family.

Mike DB5KN, Chairman, Roesrath Gigahertz Meeting

EA1DKV SK

Sad to say that Pepe EA1DKV died on January 24th. It was a



massive heart attack. It is a great loss for the ham community in northwest Spain. He was the EA1VHF beacon keeper and builder, Dubus subscriber, VHF EA pioneer and a great person.
73, Máximo - EA1DDO

F3VS SK

F3VS Claude became a SK on January 1st 2011 after a long fight against cancer. His super strong 2m EME signal will be missed! The 2011 REF/DUBUS EME CW contest will be dedicated to F3VS.

DL3XW SK

Hans Heinrich Götting, DL3XW, became a SK on January 14th 2011. He was 86 years old. Hans was one of the 2m EME pioneers in Germany in the 1960's and he had a big yagi array for several years. Together with DJ1SL he was the owner of the Götting KG and produced transceivers and amplifiers mostly for 2m. The famous HG51 PA with one 8874 tube is still in use in many shacks today.

IV3GBO accident

SV1BTR writes: My good friend Giampetro IV3GBO called me today from the clinic where he is since June 2010. At that time he had a terrible car accident, crashing with 200km/hr. For 2 weeks he was in a coma and on a wheelchair since then. Being a winner, there are reasons to believe Giampetro will be able to walk and back to radio & 2m cw EME with his 6*19LLY's and excellent signal, in 6-8 months, when he hopefully exits the specialized hospital. I wished IV3GBO the best for a good recovery. 73 Jimmy SV1BTR

New DXCC Southern Sudan

Southern Sudan will become an independent country (again) probably in June 2011 after 99% of the people voted in favour a separation from the northern part of Sudan (ST2). It is most likely that Southern Sudan will become also a separate DXCC entity again then. It has been already from 1949 to 1995 (prefix ST0) when the U.S. supported the southern groups which fought for independency.

Interesting MMIC

The new BGA7027 MMIC from NXP is a one-stage high linearity 0.5W amplifier, offered in a low-cost surface-mount package. It delivers 28 dBm output power at 1 dB gain compression (most others deliver only about 20dB) and a superior performance from 400 MHz up to 2700 MHz. The NF is about 3dB. Gain about 11dB at 2 GHz, integrated active biasing, external matching allows broad application optimization of the electrical performance, 5 V single supply operation, ESD protection at all pins. Thus this device is interesting for use on 70cm, 23cm and 13cm amateur bands. Henning Weddig, DK5LV

New from KONNI: 70 MHz Yagis

Konni Antennas offers now also low cost Yagis for the 4m band. There will be available a 4 element and a 5 element version. Contact info: KONNI-Antennen, Michelriether Str. 16, 97839 Esselbach, Germany. Phone ++49 9394/99800, Fax: / 99801, Web: www.Konni-Antennen.de

Dates

April 9: Seigy Microwave Meeting, France

April 12: EME: "Yuri's Night"

April 17: Martlesham Roundtable, U.K.

May 13 - 15: SM EME Meeting (in SM4)

June 3 - 5: Nordic VHF Meeting

Gjøvik, JP50IT, Norway. <http://arkiv.nrli.no/hamar/VUSHF2011.html>

Sept. 10 - 11: UKW-Tagung Weinheim

Deadline for next issue 2/2011: April 30th
Redaktionsschluß Ausgabe 2/2011: 30. April

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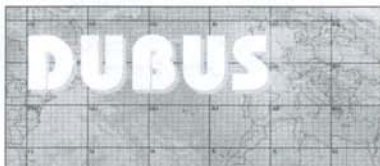
on the web – check and register at <http://portal.dubus.de/>

Beacon Lists and Toplists ONLINE - free access for all DUBUS subscribers

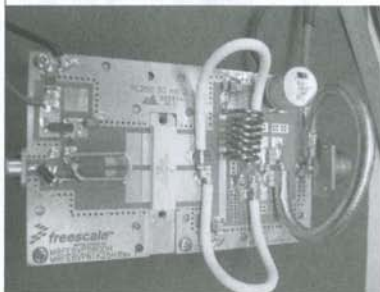
NEW! NEW! NEW!

DUBUS Book TECHNIK X

Available now!



TECHNIK X



Technique for Amateur Radio on VHF/UHF and Microwaves

The new DUBUS Technik book 10 is ready from the printer now. Size and is the same as with the previous books. On 386 pages you find the technical articles from DUBUS Magazine from the years 2009 and 2010. Some corrections and additions have been made.

The price for the book is again 25.00 Euro incl. shipment for DL and EU. Price for overseas airmail is also 25.00 Euro or 30.00 USD.

You can find the contents page on the DUBUS website at www.DUBUS.org You may order directly from DUBUS DL or via your local distributor. See also order form on page 3 in this issue.

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Requests and orders please to
funk-telegramm@t-online.de or
 DUBUS, Grützmlenweg 23, 22339 Hamburg

New Products

New from DB6NT

Transverter module MKU 432 G2

Combinations of HF transceivers and transverters provide high input sensitivity, good large signal performance, excellent phase noise values and low intermodulation distortion. Usually, the performance is much better than that of modern VHF transceivers. The high performance transverter module MKU 432 G2 is based on several decades of development and production of transverters. Its outstanding technical data makes it usable for many applications. Features:

- Low noise converter with excellent large-signal performance
- There are two separate IF connectors. They can be switched to one common IF connector.
- Switchable IF input power range
- Phase noise of the oscillator -156 dBc/Hz @ 10 kHz
- Internal temperature stabilized crystal oscillator with QH40A

Specifications:

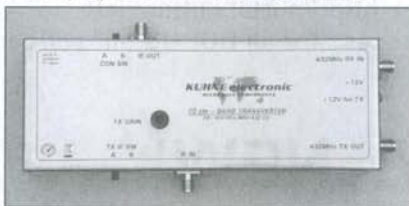
Type MKU 432 G2, VHF range 432 ... 434 MHz

IF range 28...30 MHz, IF input power 2...50 mW, adjustable /

switchable to range 60...2000 μ W, Output power min. 70 mW

RX gain typ. 25 dB, Noise figure @ 18 °C typ. 1.1 dB

Dimensions (mm) 150 x 55 x 30, Case German silver



Technically skilled customers can set up their own 432 MHz transverter system with the transverter module and one of Kuhne electronic's power amplifiers MKU PA 4047-60 HY or MKU PA 4047-120 HY. The transverter module MKU 432 G2 will work with most HF transceivers, which have a transverter output port. Due to switchable input power range and connector configuration, the module is very flexible.

MKUBPF24-SMA band pass filter for 24GHz

MKU BPF 24-SMA is a new 3 pole band pass filter in waveguide technology for image frequency rejection for 24 GHz transverters. The filter has a center frequency of 24048 MHz and a 3 dB bandwidth of typ. 250 MHz.

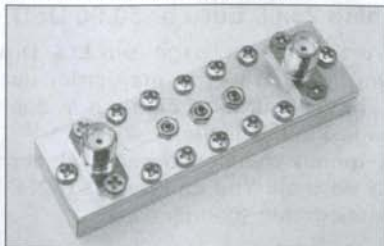


Image frequency rejection @ IF = 144 MHz typ. 18 dB

Image frequency rejection @ IF = 432 MHz typ. 48 dB

Insertion loss max. 3.0 dB, Input and output connector: SMA-

female, 50 ohms, Case: milled aluminium, dimensions (mm), 59 x 18 x 11, weight typ. 22 g

Power Amplifier MKUPA101HLK for 10 GHz Special offer – As long as stock lasts!

This new power amplifier MKU PA 101 HLK using PHEMT technology provides high linearity and high efficiency. Since the amplifier is thermally very stable and highly linear, it can be used for all operating modes. It is suitable for all analog and digital modes like SSB, CW, ATV and DATV. The amplifier features a detector output (DC voltage) for monitoring of the forward output power. Features:

- Detector output (DC voltage) for monitoring fwd output pwr
- Small mechanical dimensions
- Suitable for all analog and digital modes (SSB, CW, ATV, DATV)

Frequency range 10300 ... 10400 MHz, Input power typ. 200 mW, Output power min. 2 W (SSB/CW), Supply voltage +12 ... 14 V DC, Current consumption typ. 1.1A, connectors: SMA-female, 50 ohms, milled aluminium casing.

More information is available on our website www.db6nt.com
Kuhne electronic GmbH, Scheibenacker 3, 95180 Berg, Germany

InnovAntennas by G0KSC

A new British Yagi antenna company launches March 1st

On 1st March 2011, Justin Johnson, G0KSC launches InnovAntennas Limited to fill a niche in the global antenna market; high-quality design and construction in one package. Justin has become well known globally for his LFA (Loop Fed Array) Yagi and the marked impact it has had upon specialist areas such as EME (Earth Moon Earth) communication. He also introduced the OWL (Optimised Wideband Array) Yagi to the world with its combination of high-performance and wide bandwidth to the ham community last year. Like the many other manufacturers that have already adopted the G0KSC designs (under license), InnovAntennas will produce a wide range of these antennas in addition to the unreleased OP-DES (Opposing Phase Driven Element System) Yagi which has been specifically designed to cover individual HF bands in their entirety without the need for an ATU or antenna matching device which can badly effect the efficiency and performance of Yagi antennas.

Mechanically, InnovAntennas will aim for the top-end of the market, complementing the quality electromagnetic designs which have been created with Advanced packages such as FEKO, a specialist electromagnetic design tool which uses Particle Swarm computer Optimisation techniques and has the ability to model all aspects of the antenna including insulators, booms, buildings and even the effects people around the antenna will have.

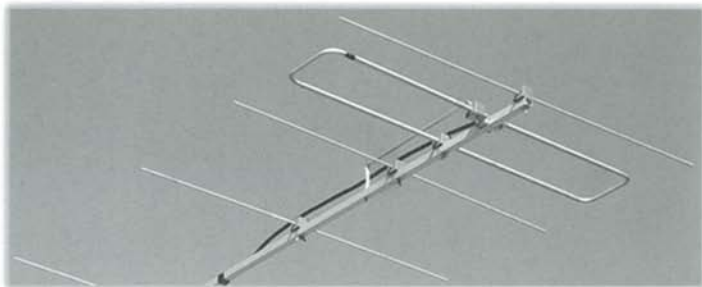
The InnovAntennas website provides in-depth information on each and every antenna in the range although the team are very happy to discuss any requirements individuals may have and will pay for the call too!

InnovAntennas can be contacted on freephone (UK only) 0800 0124 205 or ++44 800 0124 205 or via their website, www.innovantennas.com – info@innovantennas.com

InnoAntennas

Performance through Innovation

**Serious about EME or MS on VHF/UHF?
Nothing else gets close to the LFA Yagi!**



The LFA Yagi provides some of the lowest sky temperature and G/T figures ever seen. So if you work VHF/UHF weak signal modes or just suffer with high city noise that you want to get rid of, visit our website for more information today!

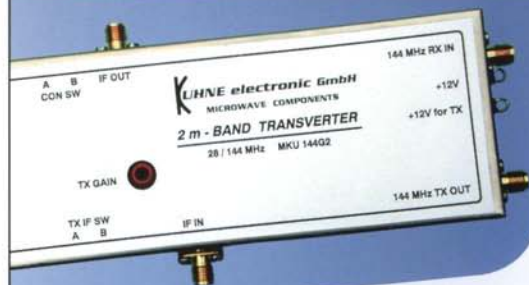
The LFA (Loop Fed Array) Yagi provides super low-noise performance. Check out our website for more information on this and other designs

Or Call Us: +44 800 0124 205

We are very happy to hear from you and will spend as much time as you need!

British Innovation, Design & Build - www.innovantennas.com

Transverter modules for 2 m and 70 cm



A well-engineered design, the use of latest components and several decades of experience in development, manufacturing and application form the basis of our transverter modules for 2 m and 70 cm. By means of switchable configurations these modules can be operated with most of the current HF transceivers with transverter output jack.

Features:

- Low noise figure and large signal performance
- Separate IF connectors (can be switched to one common connector)
- IF input power switchable
- Temperature compensated crystal oscillator with precision crystal heater (40 °C)

Type	MKU 144 G2	MKU 432 G2
VHF range	144 ... 146 MHz	432 ... 434 MHz
IF range	28 ... 30 MHz	28 ... 30 MHz
IF input power	2 ... 50 mW, adjustable / switchable to range 60 ... 2000 μ W	2 ... 50 mW, adjustable / switchable to range 60 ... 2000 μ W
Output power	min. 100 mW	min. 70 mW
RX gain	typ. 25 dB	typ. 25 dB
Noise figure @ 18 °C	typ. 0.9 dB NF	typ. 1.1 dB NF



It is possible to set up a 144 MHz or 432 MHz transverter system with the transverter module and one of the power amplifiers

MKU PA 1317-60 HY (60 W out) or
MKU PA 1317-120 HY (120 W out)

MKU PA 4047-60 HY (60 W out) or
MKU PA 4047-120 HY (120 W out)

KUHNE electronic

MICROWAVE COMPONENTS

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For further information please visit our website

www.db6nt.com