4-Square Antenna in theory and practice

Observations and experiences
of
DF6QV

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Preface

A couple of years ago I became interested in 4-Square-antennas. It is an antenna system which, as the name suggests, consists of 4 radiators, a coupler, phasing-lines, as well as a radial or ground system. The coupler is the core of the system. It distributes the output power to the radiators with proper amplitudes and phases.

This article will firstly discuss the components of this antenna system. Afterwards the influence of ground conditions and distance between the elements is shown. Different couplers will be analyzed using the simulation program LTSpice by Linear Technologies. The results, referring to the splitting of power, phasing and bandwidth are displayed in diagrams. Finally some photos hint at the practical construction of 4-square.

I'd like to express my thanks to DJ2YA for his competent support. My further thanks go to the many Oms who have enabled me to gain these experiences.

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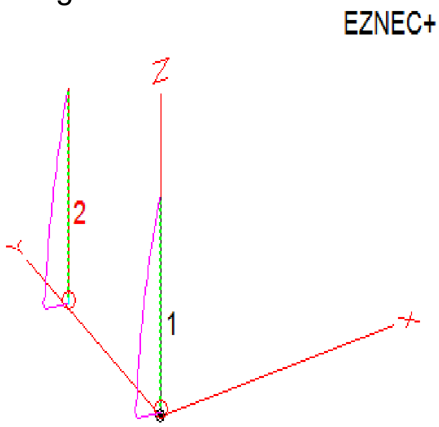
1 Antennas, Arrays, 4-Squares

A transmitting antenna couples the power generated by the transmitter to the ether. The fields generated by the transmission of power in the antenna elements overlap in the far field forming the radiation pattern of the antenna. In a receiving antenna the electromagnetic fields generate the input power for the receiver. Often transmitting and receiving antennas are the same. The current, or rather the current segments in the antenna produce fields. Their superposition results in the radiation pattern of the antenna. EZNEC and MMANA are two antenna modeling programs with a NEC2 core that use this principle.
The gain of an antenna is a measure of the energy concentration over the spacial angle. An isotropic antenna is a theoretical reference antenna that radiates uniformly into all directions. This cannot be achieved in the real world. Gains over isotropic antennas are specified as dBi. The half-wave dipole antenna is often used as a further reference. Gains over the half-wave dipole are specified as dBd. The specification of antennas should always refer to these standard antennas in order to ensure the comparability of different models.
Warning: In their prospectuses some companies include the additional gain of ground reflections to promote their antennas.
Nowadays antennas can be planned on the computer before you start building them. With the simulation of antennas you can let your scientific curiosity ramble freely. However, you should always consider whether it is possible to implement the model in the real world.
It should be remembered that our amateur bands have a certain bandwidth. In IARU Region 1 the 80m band has a relative bandwidth of about 8 % and about 13% in IARU Region 2. Therefore you must ask the following questions. What bandwidth have the radiators? What bandwidth has the network for phase shift generation and power distribution?
An antenna array is characterized by the fact that each array element is fed directly. Array elements can be dipoles, Yagis, ground planes, just any antenna you can think of. The directivity of an array is achieved by suitable phase and amplitude distribution in the individual array elements. Changes of phase and amplitude distribution in the elements will change the directivity of the antenna system. Thus it is possible to change beam direction without mechanical rotation of the antenna itself. This is an advantage especially on the lower bands (160m, 80m and 40m), where rotatable directional antennas are very bulky and expensive.
As the name implies a 4-square is a square with an antenna element at each corner. In the majority of cases they are verticals such as ground planes or vertical dipoles.

1.1 Simulation of a Two-Element Array

The following example is a 2-element array. It consists of two λ/4-spaced λ/4-verticals. The two radiators are fed with equal amplitudes but different phases. The data for the simulation are shown in Figures 1.1 to 1.3.



Array data for the simulation :

>	Vertical over real ground
File	card_no radials7mc.EZ
Frequency	7 MHz
Wavelength	42,8275 m
Wires	2 Wires, 60 segments
Sources	2 Sources
Loads	0 Loads
Trans Lines	0 Transmission Lines
Transformers	0 Transformers
L Networks	0 L Networks
Ground Type	Real/MININEC
Ground Descrip	1 Medium (0,005, 13)
Wire Loss	Zero
Units	Meters
Plot Type	Elevation
Azimuth Angle	90 Deg.
Step Size	1 Deg.
Ref Level	0 dBi
Alt SWR Z0	50 ohms
Desc Options	

Figure 1.1

Figure 1.2

Wire Create Edit Other

☐ Coord Entry Mode ☐ Preserve Connections ☑ Show Wire Insulation

Wires

No.	End 1				End 2				Diameter	Segs	Insulation	
	X (m)	Y (m)	Z (m)	Conn	X (m)	Y (m)	Z (m)	Conn	(mm)		Diel C	Thk (mm)
1	0	0	0,2		0	0	10,3		40	30	1	0
2	0	10,3	0,2		0	10,3	10,3		40	30	1	0

Figure 1.3

In the screenshots below, the lobes are shown at different phase assignments, the black curves represent the lobes at 0 degrees phase difference. The azimuth plots show an elevation angle of 25 degrees. AZ30 is the plot for 30 degree phase difference, AZ60 for 60 degrees etc. This applies analogously for the elevation plots. The radiation patterns calculated with EZNEC are shown in Figures 1.4 and 1.5 for phase differences of 30, 60, 90, 120 and 180 degrees in the feed currents. Figs. 1.6 and 1.7 show the most interesting range of 75-135 degrees at intervals of 15 degrees.

1.1.1 Azimuth-Plots

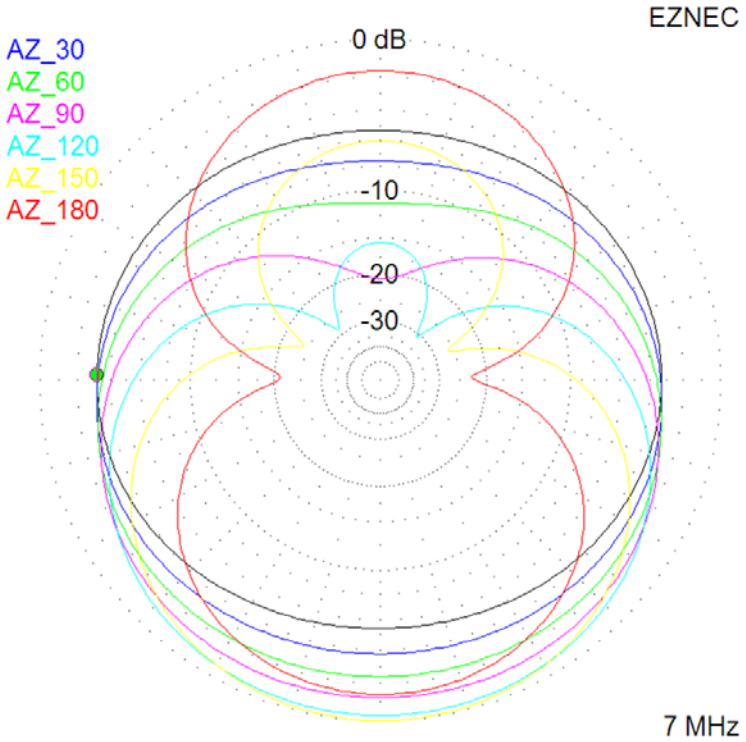


Figure 1.4

1.1.2 Elevation-Plots

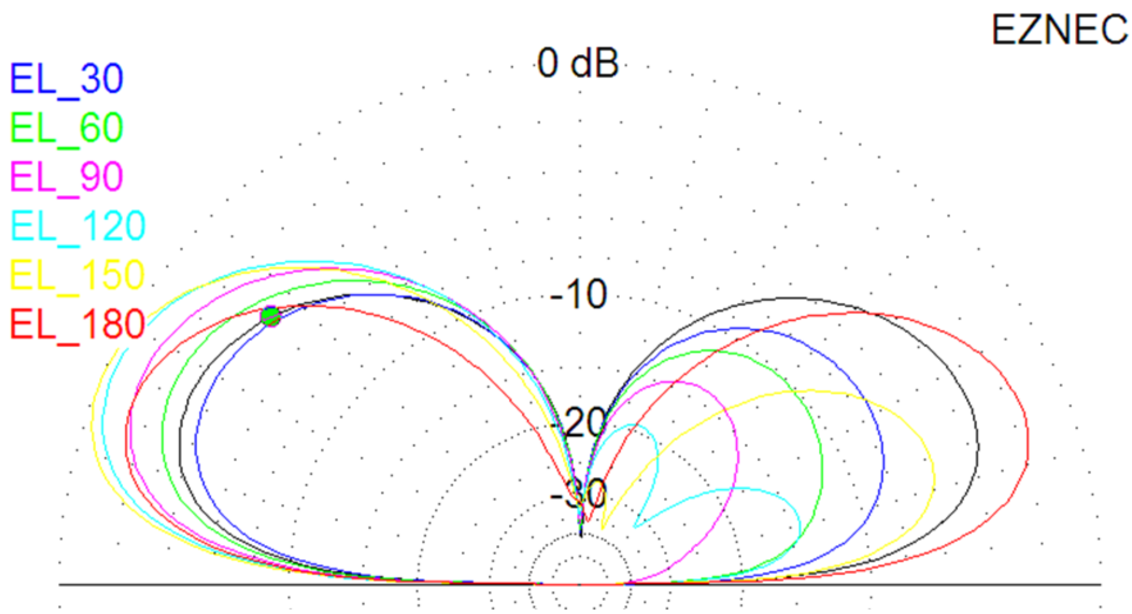


Figure 1.5

1.1.3 Plots with 75 -135 Degrees Phase Difference

The 2-element array has been simulated with an element spacing of λ/4 with the same amplitude and phase differences between 75 to 135 degrees.

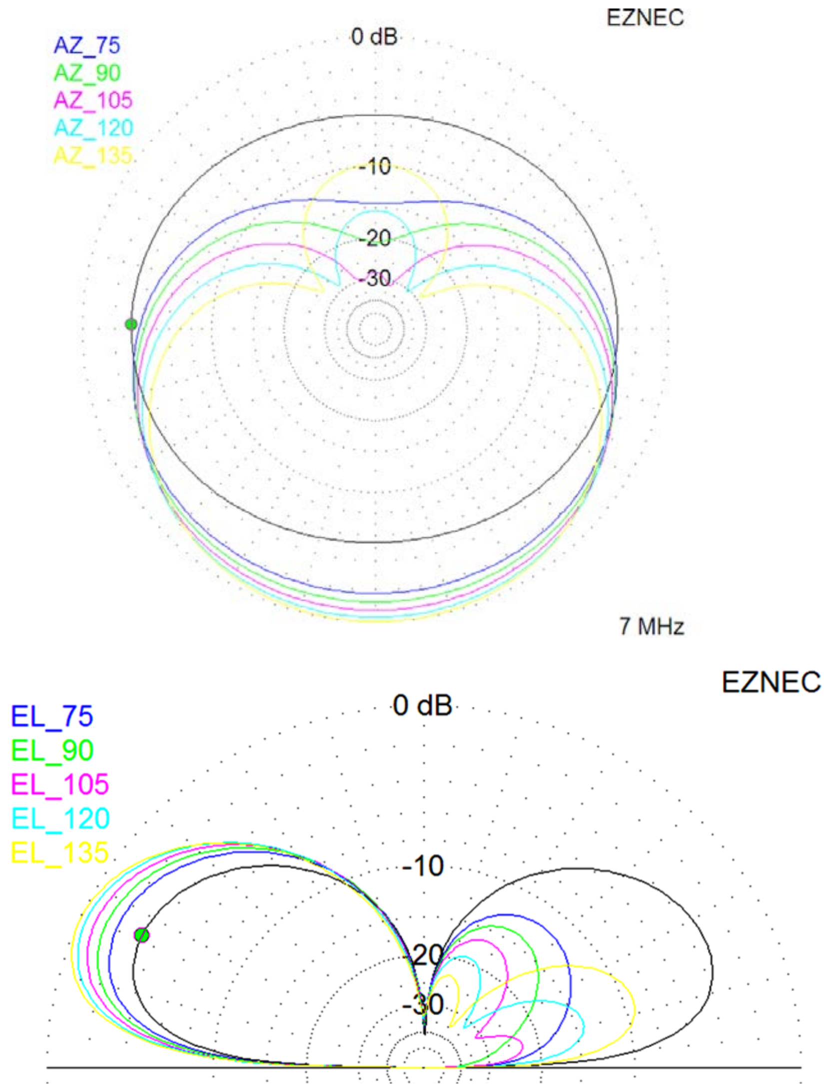


Figure 1.7

The forward gain of the 2-element array is slightly less than 3 dB (see Figure 1.8). An attractive front/back-ratio is achieved with an element spacing of λ/4 and a phase difference between 90 and 120 degrees. The gain of a λ/4 high dipole above average ground is about 7 dBi straight upward and 3.8 dBi at 28 degrees elevation. For successful DX traffic low elevation angle radiation is crucial. Interfering local signals from high elevation angles cover the DX-signal and are therefore undesirable. "Quantitative" gain figures do not count as much as the gain in "quality". This raises the question at which angle the antenna radiates best and at which angles it has only small gain or even zeroes (cf. Figures 1.8 and 1.9).

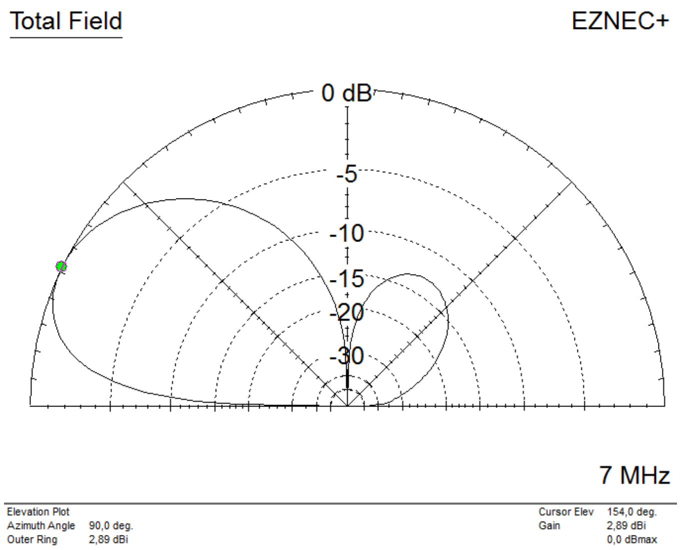


Figure 1.8

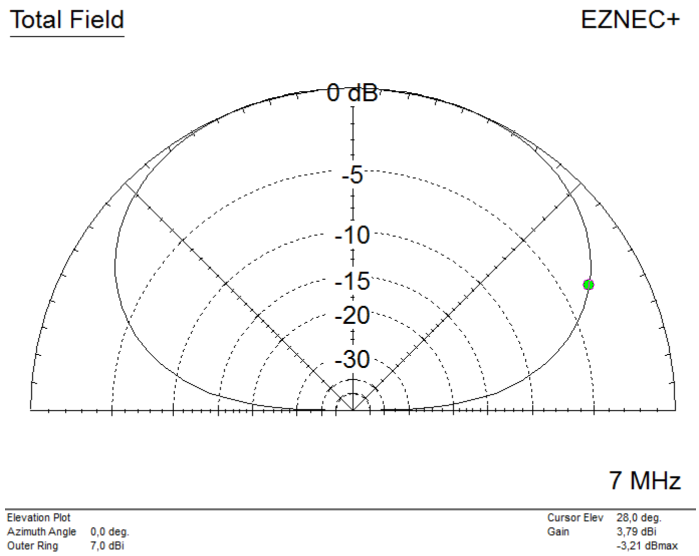


Figure 1.9

1.2 First Attempts of Array Antennas for the Low Bands

1.2.1 A Switchable Four-Element 80m Phased Array

In QST, March 1965, pp. 48 - 52, W1HKK describes an array in an article entitled "A Switchable Four-Element 80m Phased Array ". The required phase angles are generated by coaxial cables of appropriate length, impedance matching is done by λ/4-transformers. The principle is presented in Fig. 1.10.
Fig. 1.11 shows the "harness" that is required for the proper phases and impedances in the array. The figures are taken from QST, March 1965.

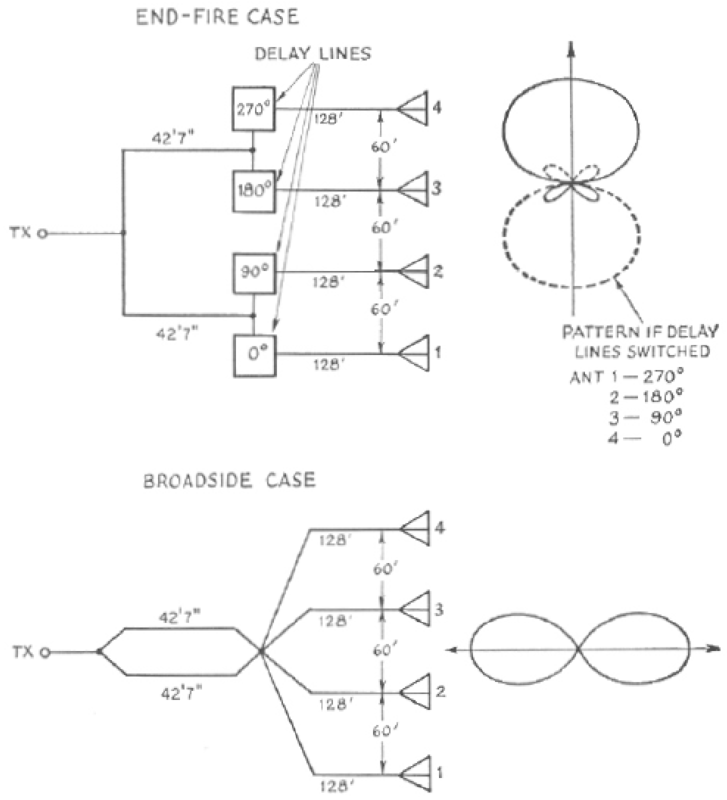


Fig. 1—Broadside and end-fire connections of the four ¼-wave antennas, with resulting patterns.

Figure 1.10

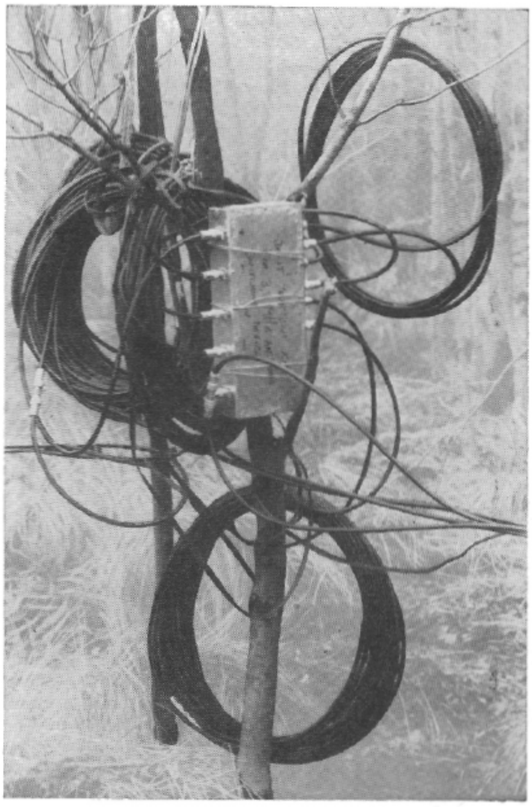
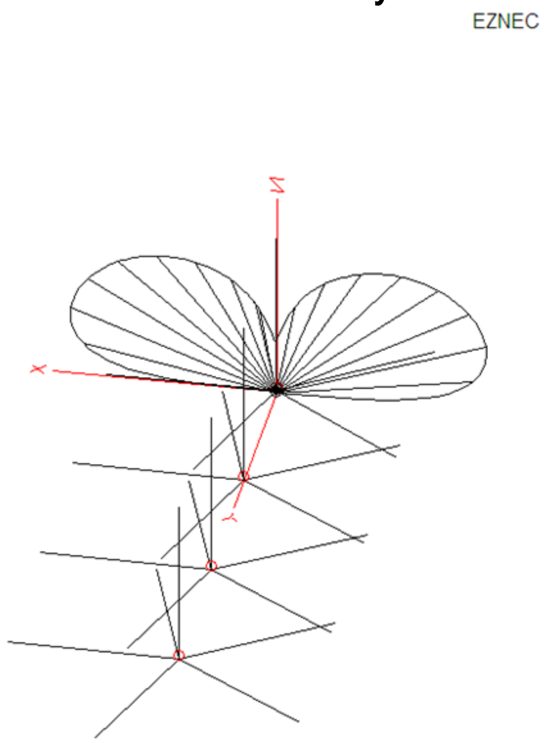


Fig. 5—The relay housing and coiled-up phasing lines.

Figure 1.11

The array does not provide for switching the length of the cables for CW or SSB. If you cut the length of the λ/4-lines for the center of the band, the error is about 4 % and 7% at the band edges. Looking at the Smith chart the small error is acceptable.

1.2.2 Simulation of the Array of W1HKK from 1965



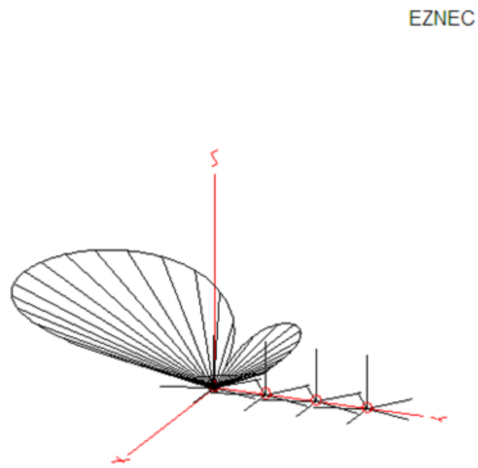


Figure 1.12 Broadside Pattern, Figure 1.13 End-Fire Pattern

The simulation has been run with 5 radials close above the ground. The patterns depend on various parameters, such as the selection of the ground type, the version of the simulation software, etc. They illustrate only the principle of the array.

1.3 4-Square Array with Ground Plane Elements

Since most 4-squares are built with ground planes, our focus is on the realization of different ground plane variants. There are several ways to make a radial system for ground plane antennas:
In a ground radial system with a number of non-resonant radials in (or on) the ground the feed impedance depends on the number and length of the radials. The more radials the lower earth losses. Above perfect ground the base impedance of a quarter wave radiator is close to 36 ohms.
Ground planes can also have elevated radials that must be resonant. The higher above ground they are, the lower are ground losses. A good compromise is achieved if you keep them at a height of about one meter in a 40m system, two meters in an

80m system and four meters in a 160m system. If two or more identical radials are arranged symmetrically, their radiation is cancelled and only the radiation of the vertical radiator is effective. Radials form a capacity to ground. The higher radials are above ground, the smaller the capacity and thus the currents flowing back through lossy earth.
Since the radiators of a 4-square are spaced approximately λ/4, there may be interactions between the radials of the individual ground planes. To avoid that it is preferable to string just one radial diagonally outward. This radial can be used to adjust the resonant frequency of the antenna. Because this radial has no symmetrical counterpart, it will cause some high-angle radiation. Such a 4-square antenna is suitable for both, local as well as DX-contacts.
A GP with one elevated radial can be regarded as a rectangular dipole. The base impedance is about 50 ohms with a full-size vertical element. If you want to avoid high-angle radiation, two or more radials must be arranged symmetrically or a Folded Counter Poise (FCP) can be employed. In the Austrian amateur radio magazine QSP a proposal has been published in which high-angle radiation is cancelled by clever folding of the radials. It should be considered that different ground systems cause different base impedances that must be matched to 50 ohms.
In QSP, 10/2009, pp. 53, OE3REB has written an excellent article on the topic of radials entitled. "120 to 4 ..."
http://oe9.oevsv.at/export/oevsv/download/QSPArchiv/2009/QSP200910.pdf
In a series of QEX articles N6LF has intensively treated the subject of radials:
http://www.antennasbyn6lf.com/2009/12/series-of-qex-articles-on-ground-system-experiments.html
K2AV has described a folded radial system at:
http://www.w0uce.net/K2AVantennas.html

1.3.1 The Principle of a 4-Square Antenna with a Hybrid Coupler

A 3 dB-hybrid coupler divides the input power into two "identical" parts with a phase difference of 90 degrees. Figs. 1.15 and 1.16 show two possible 4-square configurations. In the table below it can be seen that in the diamond structure two elements have the same phase distribution while the phases in the two other elements differ by plus and minus 90 degrees. This can be achieved by suitable phase shifting and a 180-degree-transformer. The base impedance of the radiators are transformed to approximately 100 ohms by λ/4-lines of the required impedance. By the parallel connection of two radiators a match to the 50 ohm port of the -3 dB-hybrid is ensured

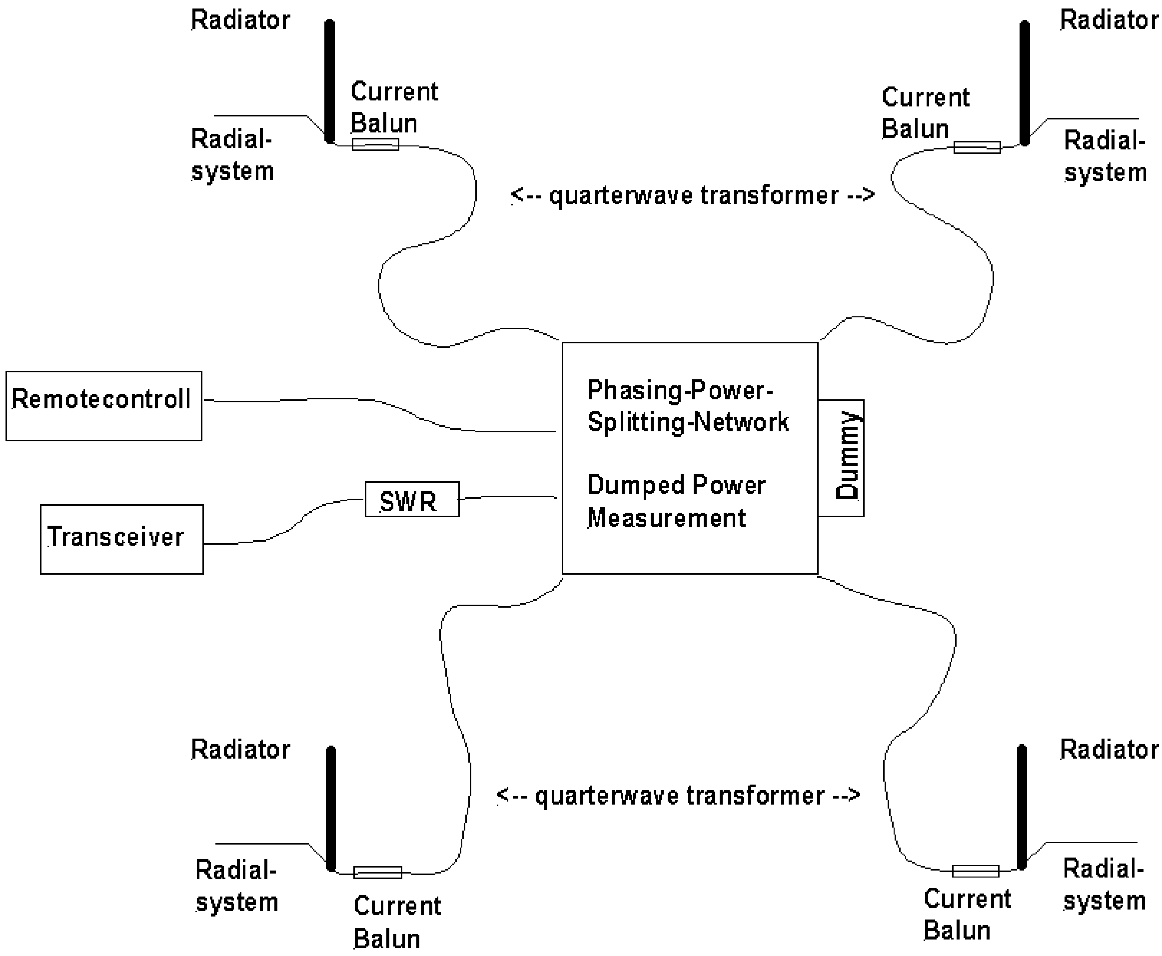


Figure 1.14 Principle of a four-square antenna system with a hybrid-coupler

1.3.2 Radiator Arrangement

Building a 4-square the following two antenna arrangements are possible, diamond shape (Fig. 1.15) and square as shown in Figure 1.16. E1 - E4 are the radiators, N, O, S and W are the four possible beam directions. The phase assignments of the radiators is shown in the table below. The square structure can be regarded as two parallel-connected 2-element arrays. Fig. 1.17 shows the respective radiation patterns.

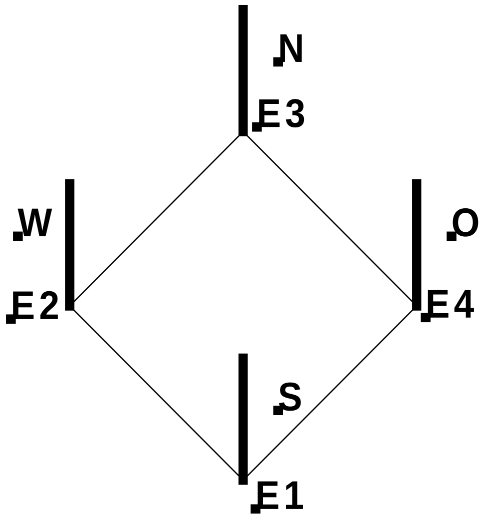


Figure 1.16 diamond shape

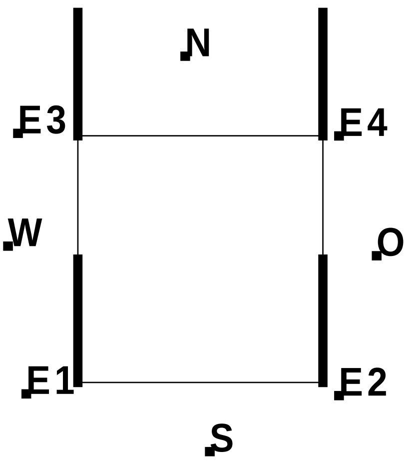


Figure 1.16 square

	Diamond			
	North	East	South	West
E1	+90	0	-90	0
E2	0	+90	0	-90
E3	-90	0	+90	0
E4	0	-90	0	+90

Table 1.1

	Quadrat			
	North	East	South	West
E1	+90	+90	0	0
E2	+90	0	0	+90
E3	0	+90	+90	0
E4	0	0	+90	+90

Table 1.2

In tables 1.1 and 1.2 the phase assignments for the elements for the different beam directions are listed. The amplitudes of the source are the same for all four elements.

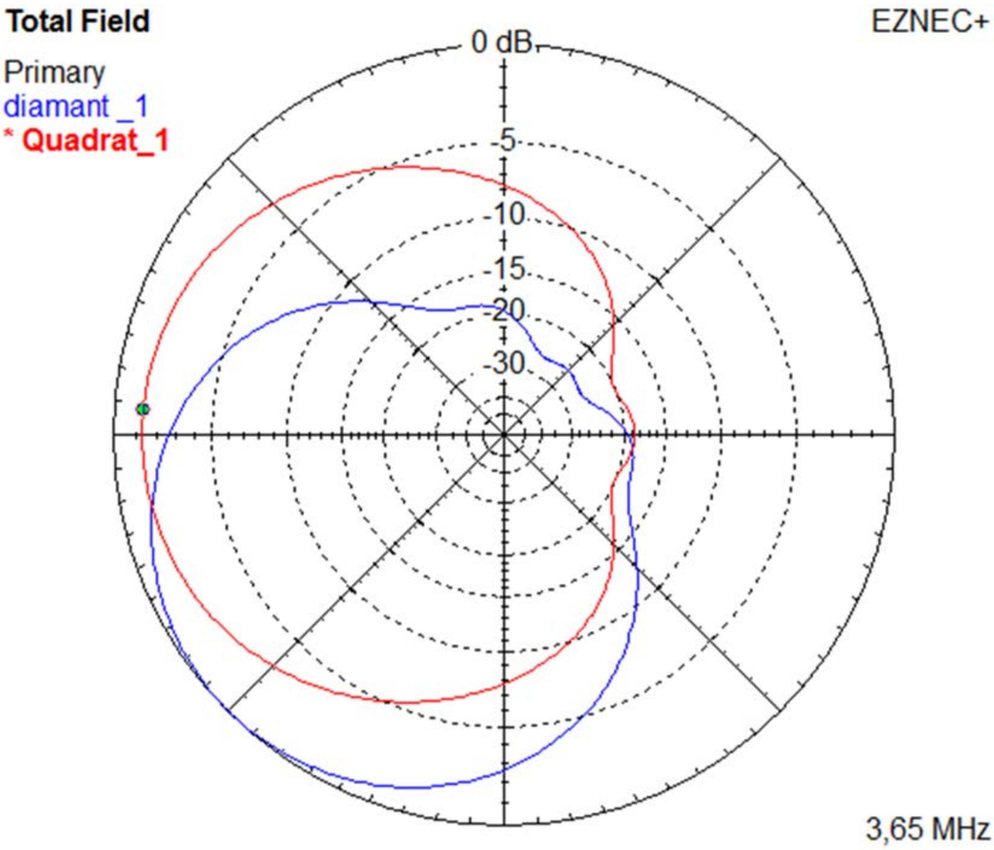


Figure 1.17 The radiation maxima differ by 45 degrees

Looking at the azimuth plots of the two arrangements in the diagram above, you can see that the diamond structure has about 1.2 dB more forward gain. At the maximum of the square structure the diamond has about 1.7 dB less gain. Therefore, a combination of both arrangements is worthwhile. This calls for a 4-square for 8 beam directions.
Entitled "A Four-Square with Eight Directions of Fire", http://ncjweb.com/bonus-content/k3lc4squarea.pdf, Al Christman , K3LC, presents a description of such an arrangement in the WEB.

1.3.3 Element Spacing in a 4-Square

An EZNEC analysis of different spacings focusses the example of an 80m 4-square with one elevated radial 2m high. The element spacing is varied in steps of 0.02 λ between 0.20 λ and 0.30 λ.

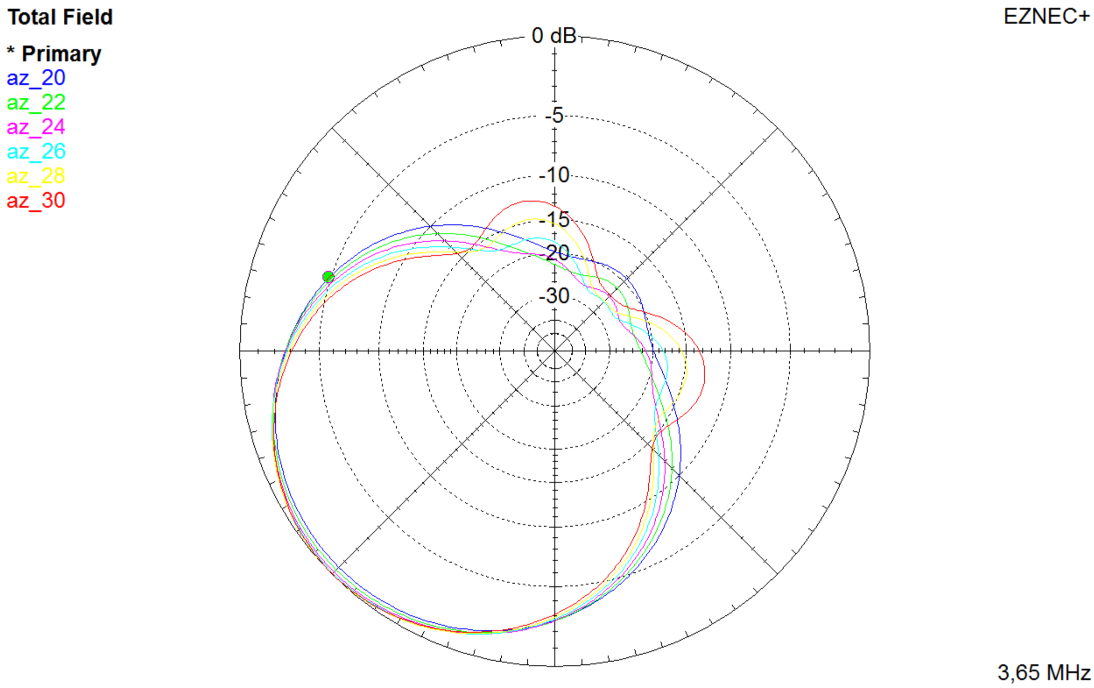


Figure 1.18

In Figure 1.18 the azimuth patterns at an elevation angle of 22 degrees over average ground are calculated with EZNEC as a function of element spacing.

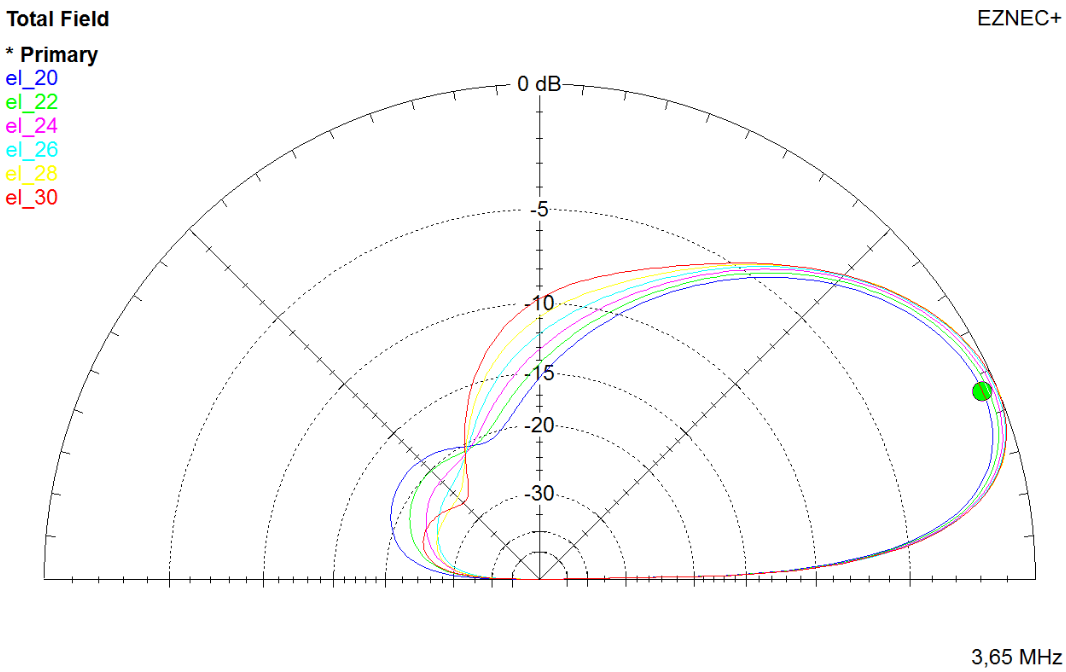


Figure 1.19

In Figure 1.19 the corresponding elevation patterns over average ground are calculated as a function of element spacing.

At greater spacing there is a minimal increase in forward gain; the rear lobes also increase somewhat and the -3 dB beam width is smaller.

The table shows the -3 dB beam-width listed as a function of element spacing in λ. These data were obtained when modeling the 80m 4-Square for TS7N with EZNEC.

Spacing / λ	Beam width / Deg.
0,20	105,6
0,22	101,2
0,24	97,6
0,26	94,4
0,28	91,6
0,30	89,0

Table 1.3

The table shows that λ/4-spacing is not really a must!

4-squares with optimized parameters (amplitude, phase shift and height) for maximal gain have a smaller beam width.
Note: Gain means concentrating energy in a spacial angle!

The higher the gain the smaller the -3 dB beam-width. For larger spacings there is probably no way around Al Christman`s 8-direction solution.

1.3.4 Ground Influence

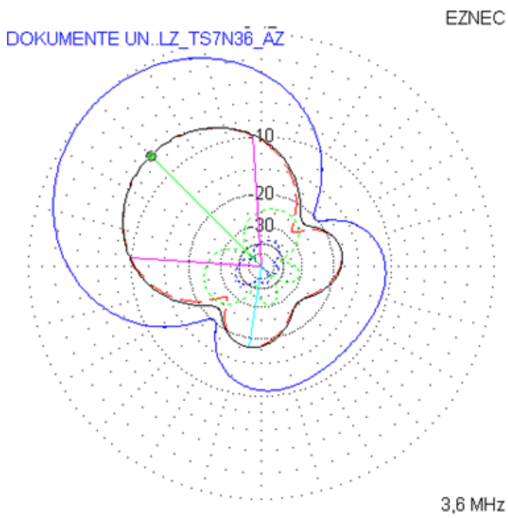


Figure 1.20

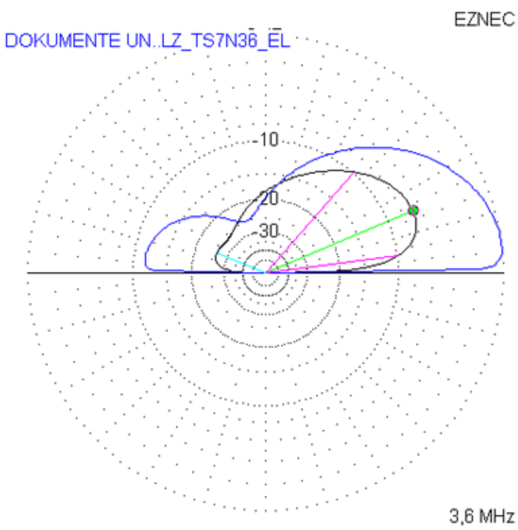


Figure 1.21

The TS7N antenna was modeled over average ground with a conductance of 0.005 and a permitivity of 13 and also above salt water (S = 5, ε_r = 80). Over salt water the gain is increased by about 6 dB, while the optimal elevation angle decreases from about 23 degrees to about 6 degrees. This is illustrated in Figures 1.20 and 1.21. With a low angle of elevation and less hops the attenuation between transmitter and receiver decreases. Reflections on salt water are relatively low loss, whereas

reflections on dry ground suffer from considerable loss. More details on propagation behavior offer the following links:
http://www.astrosurf.com/luxorion/qsl-hf-tutorial-nm7m6.htm
http://www.qsl.net/g3yrc/hf % 20propagation/hf % 20propagation.htm

2 Variants of 90-Degree-Couplers

For the operation of a 4-square the input power must be split into four equal parts. Two radiators have 0 degree phase difference, the two others have +90 and -90 degrees with respect to the other two. With 3-dB hybrid couplers you can meet this requirement almost perfectly. Other solutions of power splitting and phase assignment can be found in ON4UN`s " Low Band DXing".
Below I have examined three variants of hybrid couplers at 3.65 MHz. A fairly simple variant with coaxial delay lines, as well as two others with 90-degree 3-dB couplers are analyzed with respect to their specific properties.
1) simple 2-element array (Type A)
2) 3-dB hybrid coupler (often called Branch Line Coupler)
   a) implementation with lines (Type B)
   b) lines replaced by discrete components (Type C)
3) coupler by Reed Fisher
   a) (Type D)
   b) broadband version (type E)
The simulations were carried out for 3.65 MHz using "SwitcherCAD" or "LTSpice" , a freeware from Linear Technologies. Unless specifically indicated, loss-free components were used in the simulation.

2.1 Simple 2-Element Array (Type A)

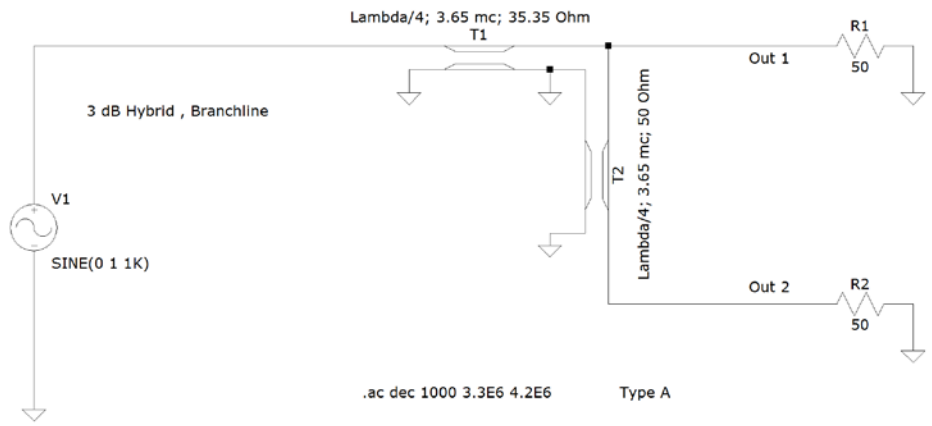


Figure 2.1

A fairly easy way to feed two radiators with a phase difference of 90 degrees is shown in Figure 2.1. The two outputs OUT 1 and OUT 2 are connected in parallel through a 50 ohm λ/4-line (T2). At this point, the system has 25 ohms. This impedance is transformed to 50 ohms by another quarter-wave transformer (T1) yielding an impedance of 35 ohms. Two parallel 75 Ohm λ/4-lines will do that quite well. Antennas only have a real resistance at resonance and that should match the system. With a mismatch, there will be reflections from OUT 1 and OUT 2 and the bounced energy will be reflected into the respective radiator and the input. This has a negative effect on the radiation characteristics of the array. Mutual coupling of the two radiators has not been taken into account.

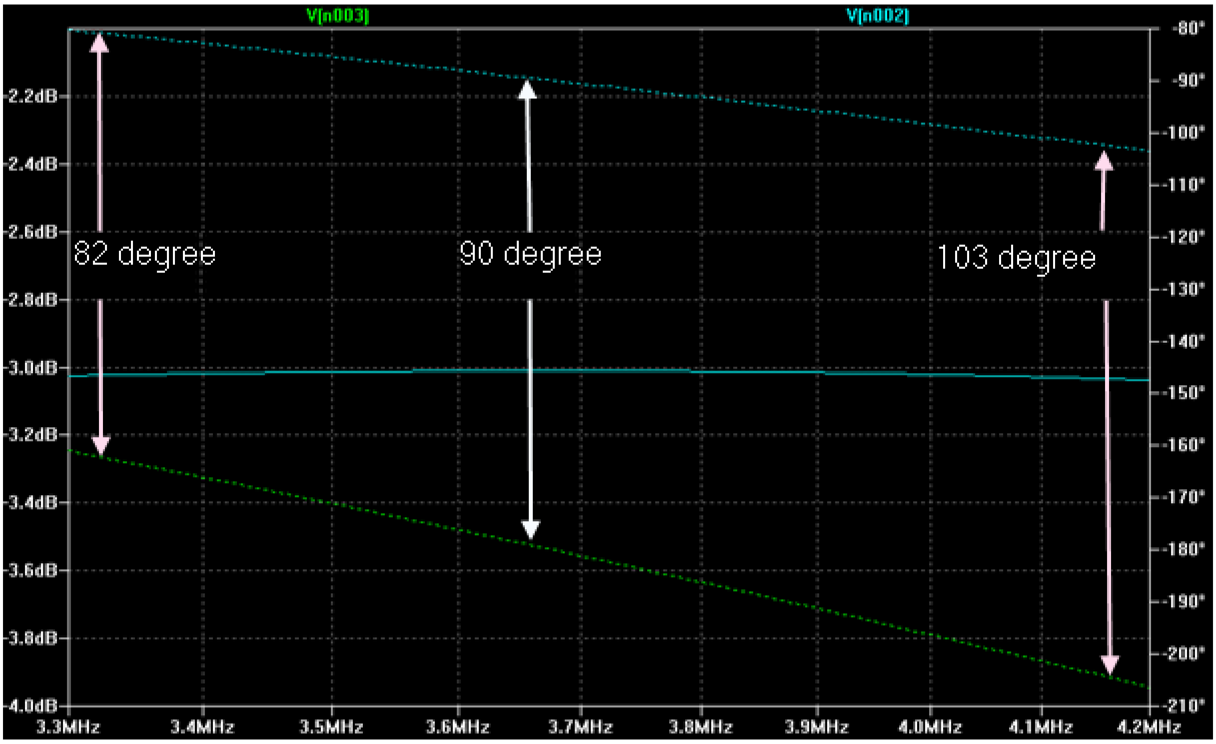


Figure 2.2

In Fig. 2.2 the amplitudes of the two outputs are shown as a solid curve on the -3 dB line. Within the frequency range of 3.3 to 4.2 MHz they are almost independent of frequency. The phase difference of the two outputs, shown by dotted lines, is frequency dependent. On the design frequency of 3.65 MHz the graph shows 90 degrees, while the phase differences at the edges of the frequency range are 82 and 103 degrees.

2.2 3 dB Hybrid Coupler (often called branch line coupler)

2.2.1 Implementation with Lines (Type B)

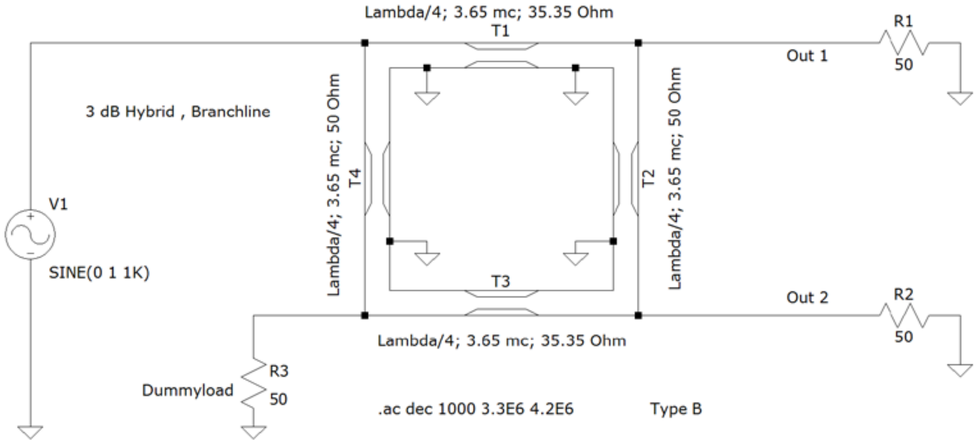


Figure 2.3

In a hybrid coupler the power splitting and the generation of the phase shift for the outputs are similar as in Type A. If reflections appear at the outputs, the reflected waves cancel at the other output. For example, power reflected at OUT 1 runs on lines T1 + T4 + T3 (270 degrees) and T2 (90 degrees) to OUT 2 in equal parts. Because of the phase difference of 180 degrees it is canceled there. The same applies to the other output and therefore the two ports are decoupled!
The power that is reflected from the outputs flows into the dummy load. The reflections reach the dummy load in phase and are absorbed there. Power reflected from the two outputs to the input port are of opposite phase and cancel. However, they cancel totally only if the two outputs are terminated equally (good or bad). As a consequence the input SWR is always good, even if an amount of power is reflected from the antenna ports, because it is dissipated in the dummy load.
Note: The input SWR is not a measure of the quality of the array!
The coupler is described in Hamradio, April / May 1984, in an article entitled: "The Branch Line Hybrid" (Part 1/ 2).
The coupling of the input power to the two outputs, the phase relationship of the outputs to each other, as well as the decoupling of the dummy port are simulated in the two screenshots below. They prove that this type of coupler is not perfect for covering the U.S. 80m band with its relative bandwidth of approximately 13%. The output amplitude errors are significant.

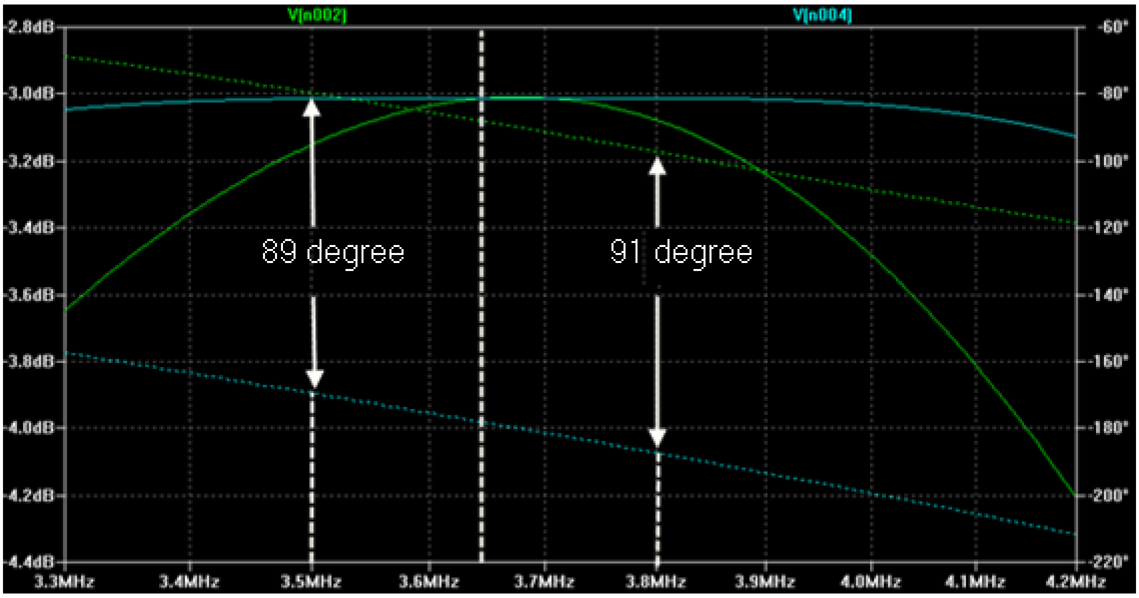


Figure 2.4

V (n002) = OUT 1 and V (n004) = OUT 2 are the plots of these outputs between 3.3 and 4.2 MHz. The amplitudes are shown as solid, phases as dashed lines. In the range of the 80m band in Region 1 the phase differences of the outputs vary between 89 and 91 degrees. The amplitude error is less than 0.15 dB.

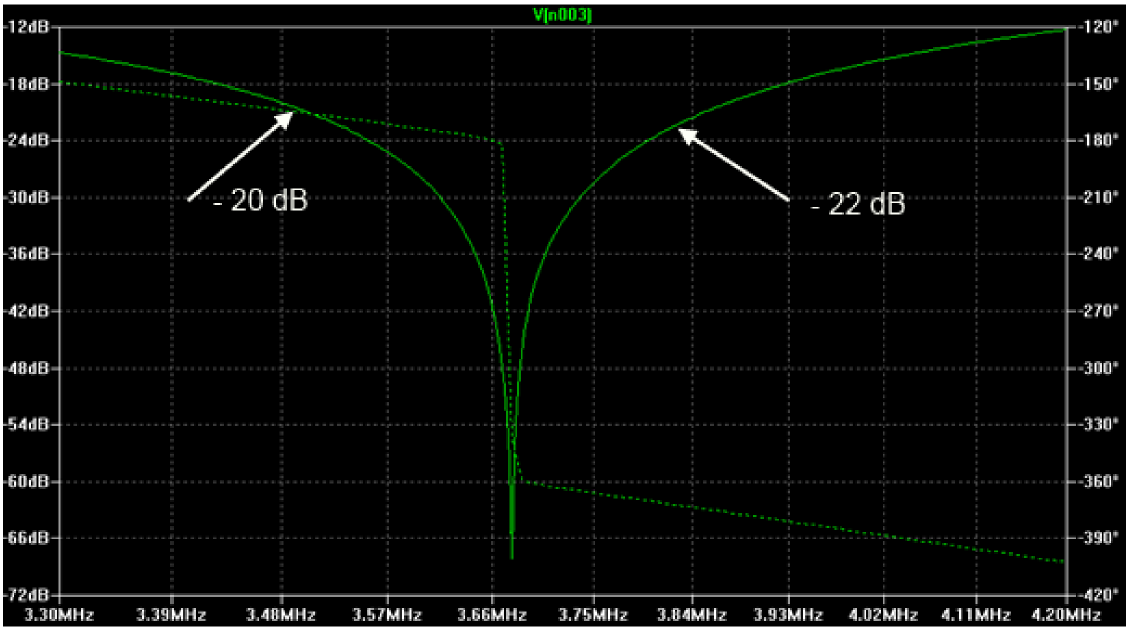


Figure 2.5

V(n003) is the plot for the dummy port. At the dummy load port, the phase is not important. The input power is distributed to ports OUT 1, OUT 2 and the dummy load.
Fig. 2.5 shows that on the design frequency (3.65 MHz) the power is distributed only to the output ports OUT 1 and OUT 2. This results in the high decoupling of

60 dB at the dummy port on the design frequency. In the full range of the 80m-band decoupling is always greater than 20 dB.

2.2.2 3 dB Hybrid Coupler with Discrete Components (Type C)

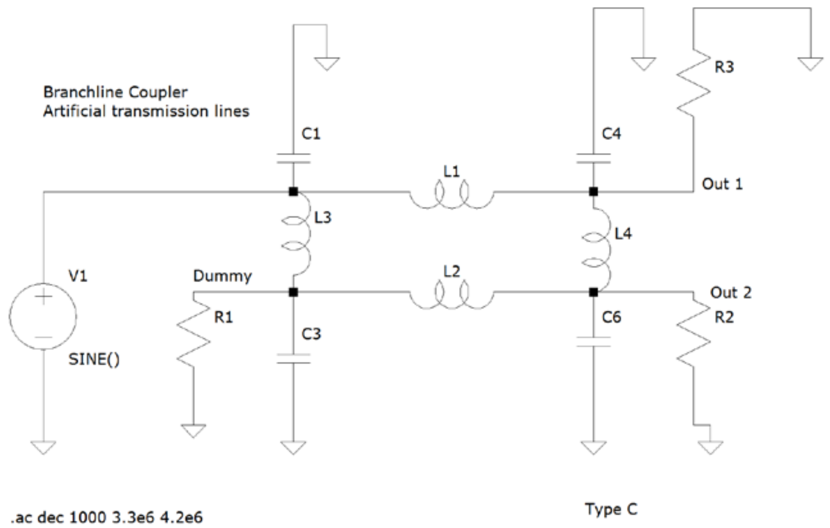


Figure 2.6

In Fig.2.6 the transformers are replaced by "artificial lines", Pi-equivalent circuits with discrete components. In the circuit diagram the four capacitors of the Pi-elements are grouped together at the corners. This type of coupler has a much narrower bandwidth than that of type B with real lines. In this way you can build a small but very narrow band coupler for the low bands, however with limited bandwidth.

2.3 The Reed Fisher Coupler

2.3.1 The Coupler according to Reed Fisher (Type D)

This type is not comparable with the structure and the function of the Branch Line Coupler with its frequency dependent coupling to the output ports. The phase difference of 90 degrees between OUT 1 and OUT 2 is nearly independent of the frequency and the decoupling of the insulation port is very high over a wide frequency range. Just as the Branch Line Coupler this coupler fulfills the requirements for a 4-square regarding power splitting and phase relationship on the design frequency:
- an equal amount of power at the two output ports
- a phase difference of 90 degrees
- good decoupling of the isolation port (dummy load)

This coupler is described in QST, January 1978, by W2CQH in an article entitled "Twisted Wire Quadrature Hybrid Directional Couplers".

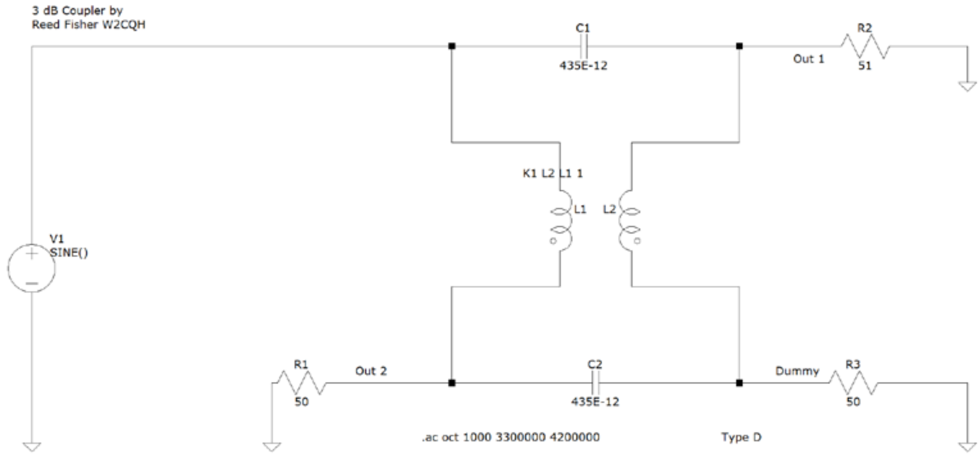


Figure 2.7

The following pictures show simulations with loss-less and loss-loaded coils for the frequency range between 3.3 MHz and 4.2 MHz. To simulate a poor quality coil a 3 ohm resistor is connected in series with a loss-less inductor.

Diagrams of the simulation with loss-less coils:

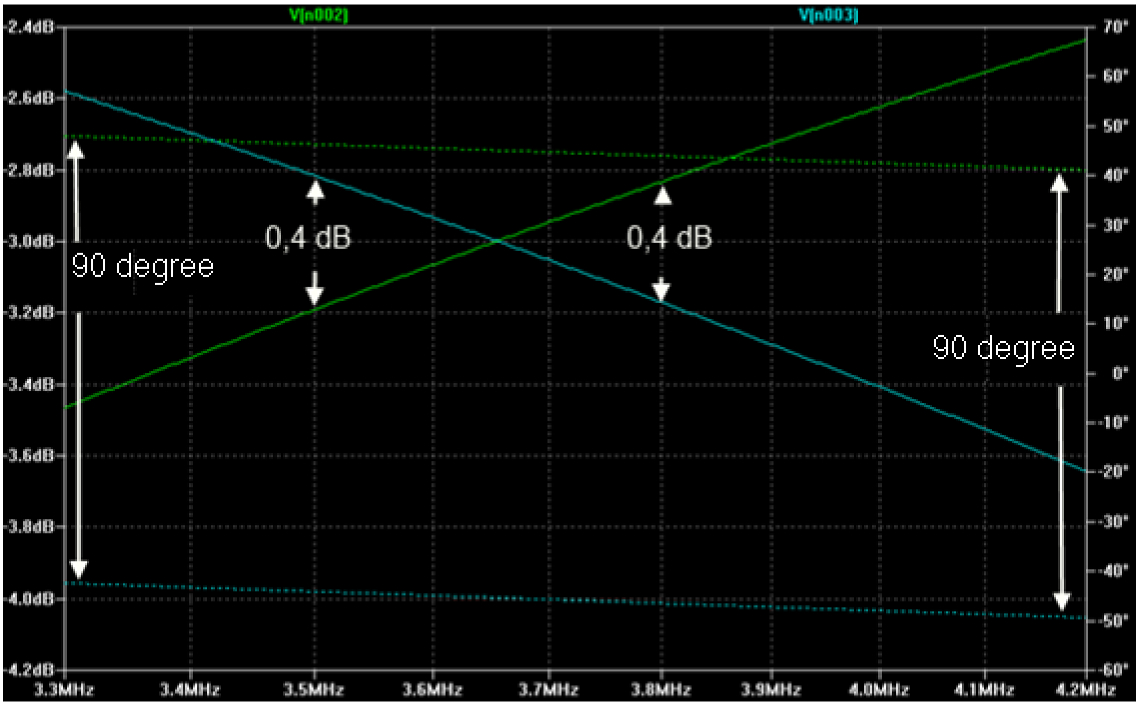


Figure 2.8

V (n002) = OUT 1 and V (n003) = OUT 2 are the plots of the outputs between 3.3 and 4.2 MHz with amplitudes as solid and phases as dashed lines. The phase difference between the outputs is constantly 90 degrees. The coupling to the two ports is frequency dependent. The maximal difference in amplitudes over the whole range of the 80m band is about 0.4 dB.

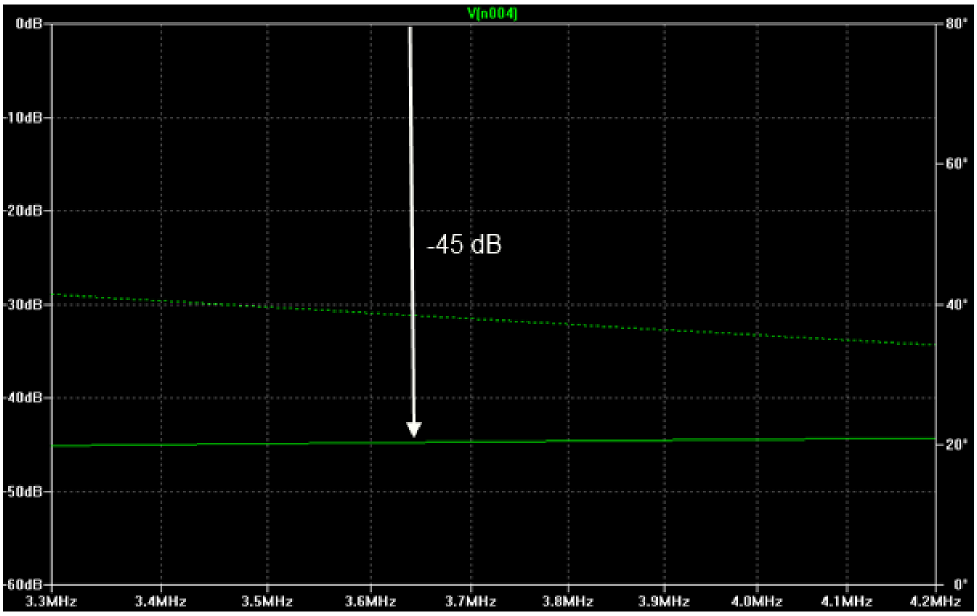


Figure 2.9

V (n004) is the plot for the dummy port. The phase angle for the power in the dummy load is not of interest. The simulation shows a decoupling of the input port of more than 40 dB (see Fig. 2.9).
For the simulation of couplers with loss a 3 ohm resistor was switched in series with the coil. The diagrams are shown in Figures 2.10 and 2.11.

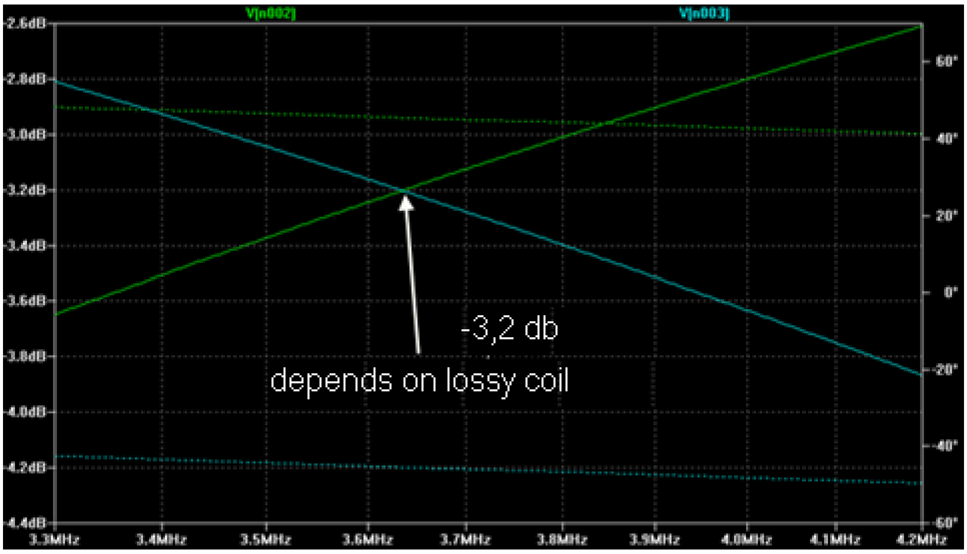


Figure 2.10

V (n002) = OUT 1 and V (n003) = OUT 2 are the plots of the outputs between 3.3 and 4.2 MHz.. On the design frequency of 3.65 MHz both outputs show 0.2 dB lower amplitudes than a coupler with loss-less coils.

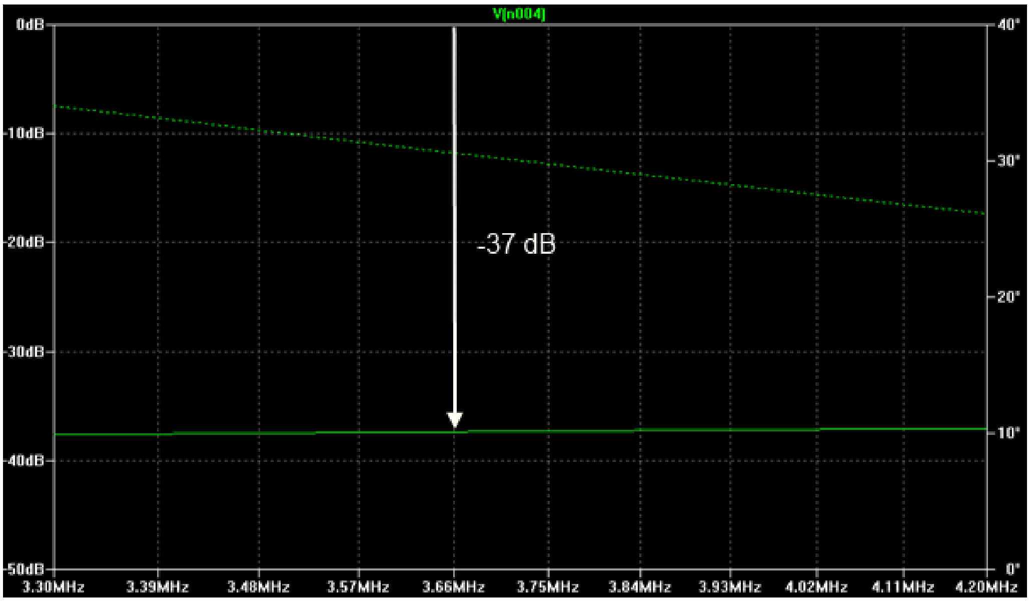


Figure 2.11

V (n004) is the plot for the dummy port. The decoupling deteriorates from -45 dB to -37 dB compared to the coupler with loss-less coils.
If you operate the coupler off the design frequency, different levels appear at the outputs. This can be desirable for special cases. An example is shown in Fig. 2.12. The image also shows that the 90 degrees phase difference remains constant over a very wide frequency range.

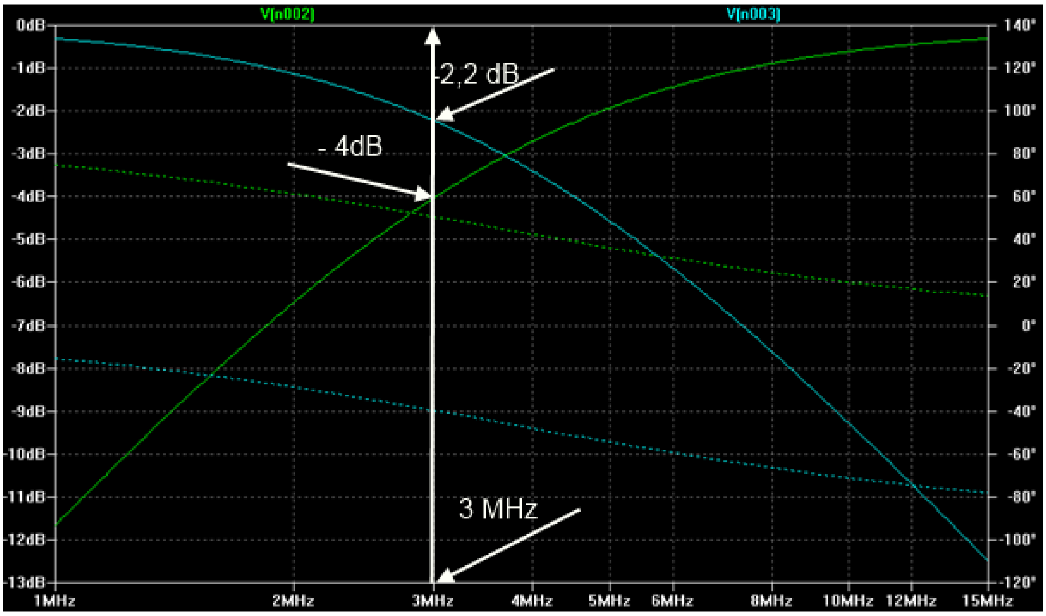


Figure 2.12

V (n002) = OUT 1 and V (n003) = OUT 2 are the plots of the amplitude outputs at these ports between 1 and 15 MHz. It is possible to obtain different power levels with a phase difference of 90 degrees as this difference remains constant over a very wide frequency range. For instance, at 3 MHz you can read a coupling of -4 dB to OUT 1 and one of -2.2 dB to OUT 2 from the graph.

2.3.2 A Broadband Version of the Reed Fisher Coupler (Type E)

In HAM Radio, June 1982, in the article "The Hybrid Coupler" by W1OOP you find an idea by W2CQH to give his coupler more bandwidth. He connects two of these couplers via a twin coax line of about 23 degrees electrical length. The circuit is shown in Fig. 2.13. The simulation shows an amplitude error of less than 0.2 dB over the frequency range from 3.1 to 4.1 MHz. The phase angle remains constant at about 85 degrees over the entire range. With these values the coupler is suitable for the entire U.S 80m band. To achieve these values the design frequency for the calculation of the components is determined empirically as about 5 MHz. The length of the linking line has an influence on the symmetry of the outputs. The component values are given in Fig. 2.13. .

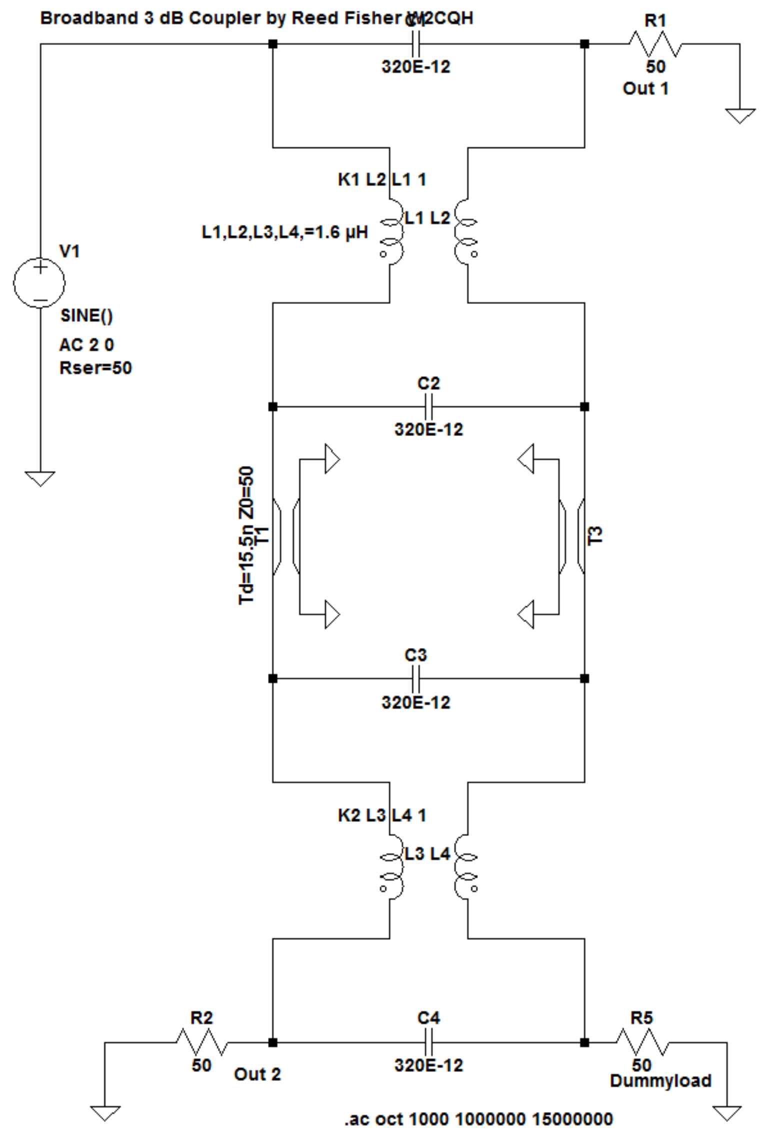


Figure 2.13 Diagram of the "broadband" version of W2CQH

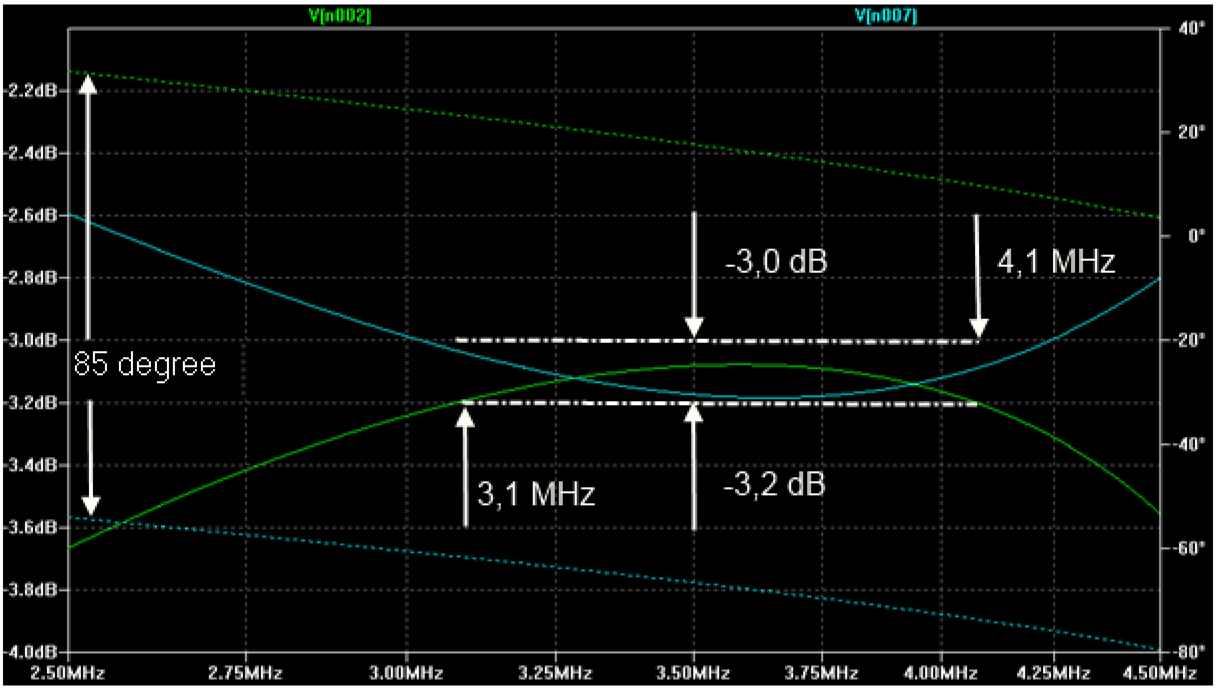


Figure 2.14

V (n002) = OUT 1 and V (n007) = OUT 2 are the plots of the outputs from 2.5 to 4.5 MHz. The phase difference between OUT 1 and OUT 2 is about 85 degrees. With the component values shown in the circuit diagram the coupling error of less than 0.2 dB between 3.1 and 4.1 MHz promises an interesting alternative. The decoupling of the dummy load port is better than 35 dB (cf. Fig. 2.15).

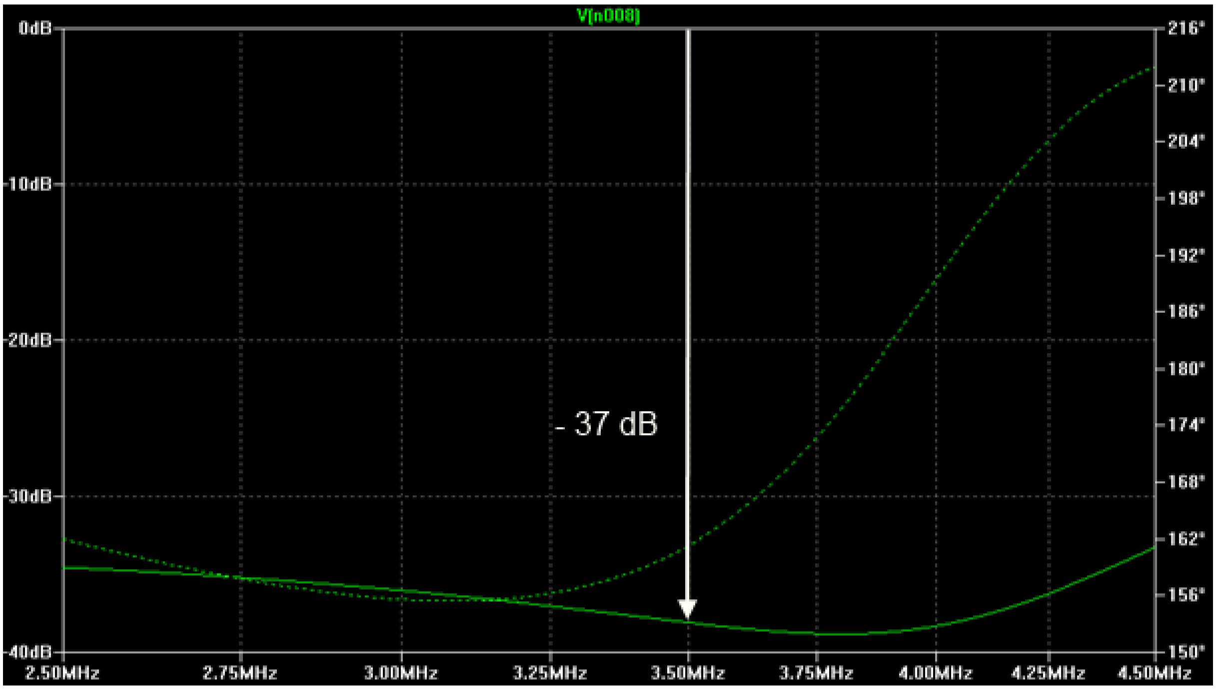


Figure 2.15

2.3.3 The Coupler as a Phase Shifter

If OUT 1 to OUT 2 are terminated with equal purely reactive loads (e.g. a non-resonant parallel circuit), the input power is totally reflected into the dummy port. The phase angle between the input signal and port "Out" is a function of the phase angle of the reflection coefficient, which is determined by the reactance at ports OUT 1 and OUT 2. For the design frequency fairly small variable phase shifts can be realized. A useful application may be feeding two parallel Beverages with a phase difference. Via a splitter one Beverage can be connected directly and the second to "Out" through the phase shifter. With variable capacitors at OUT1 and OUT 2 the phase shift between the two Beverages can be varied to steer the direction of the main lobe.

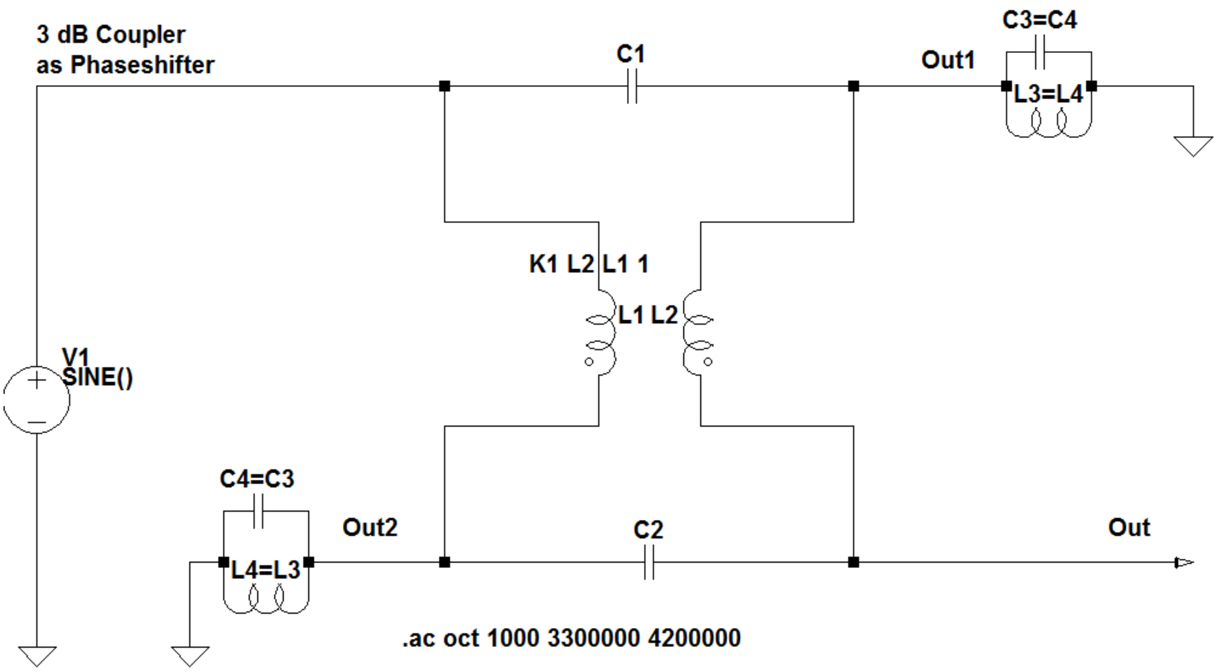


Figure 2.16 3-dB-hybrid-coupler as a phase shifter

2.4 Couplers in Practice

This chapter will discuss recorded data of two couplers that I have built.

2.4.1 Strip-Line Coupler

For my thesis in 1980 I have built and measured a 315 MHz coupler in strip-line technology of type B. Figs. 2.17 – 2.20 show the results:

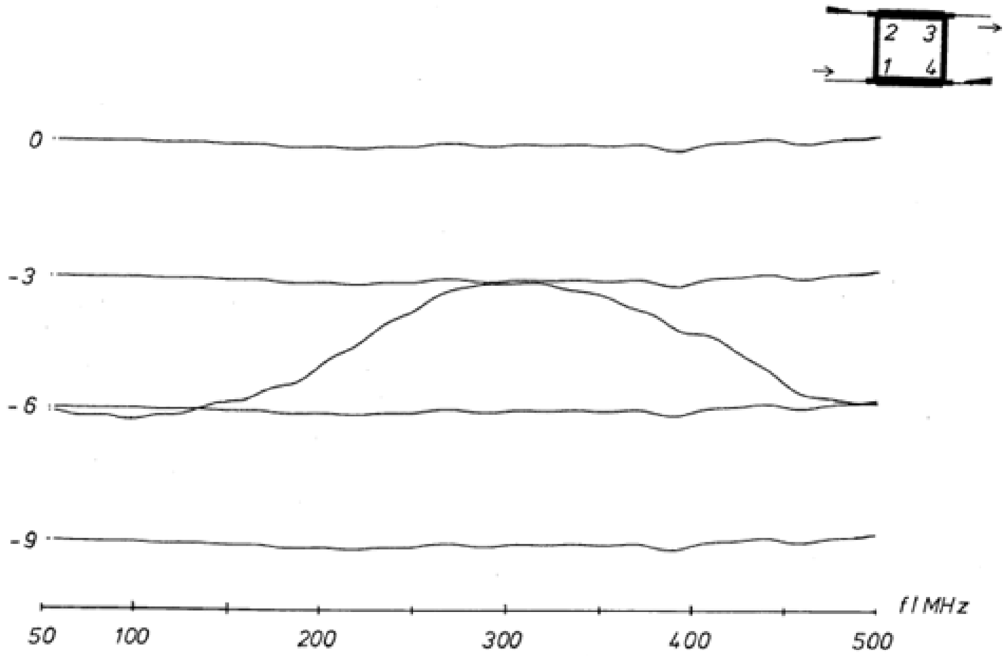


Figure 2.17 shows the coupling to Port 3

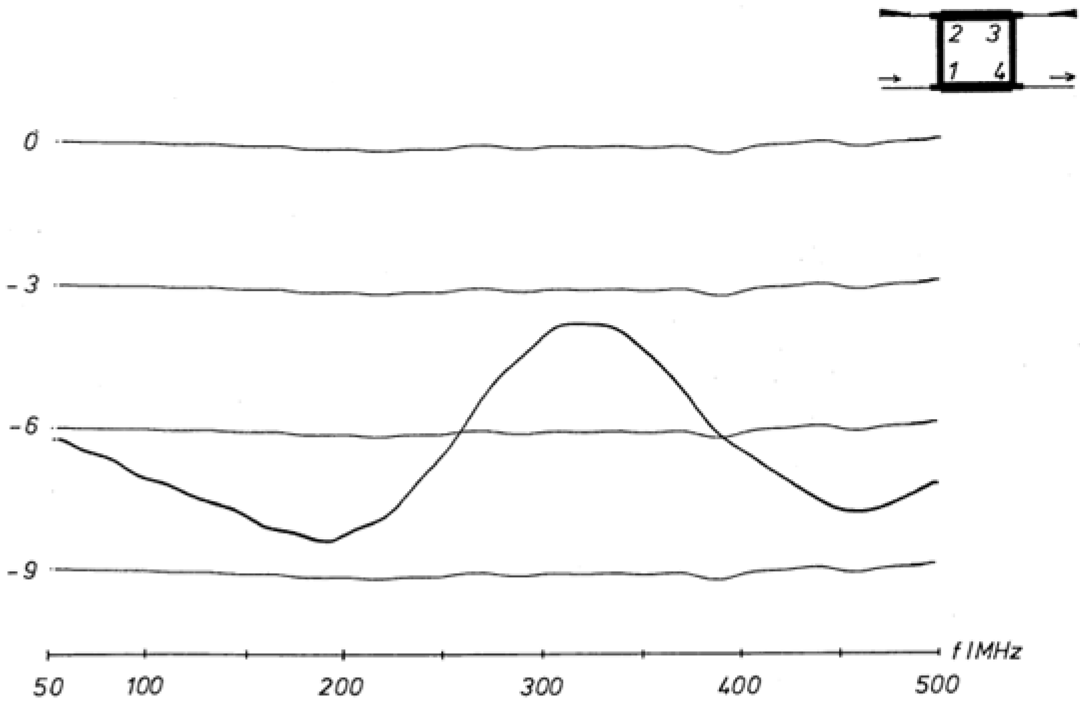


Figure 2.18 shows the coupling to Port 4

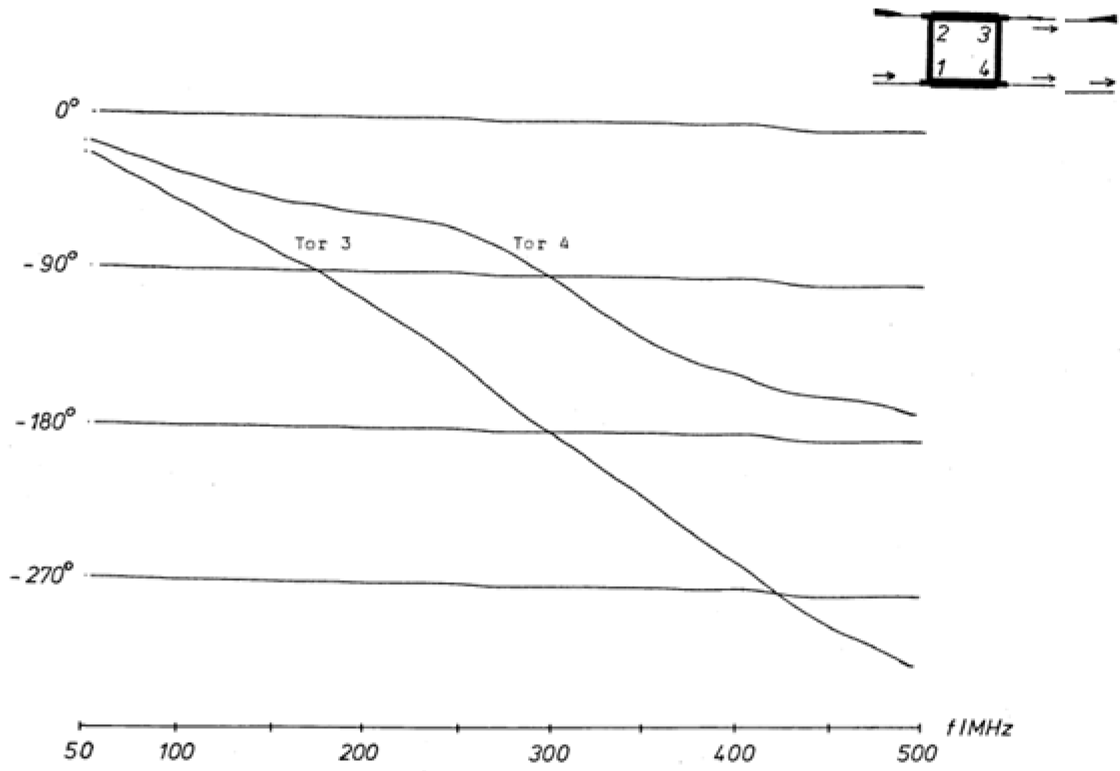


Figure 2.19 The phase difference at the design frequency is 90 degrees

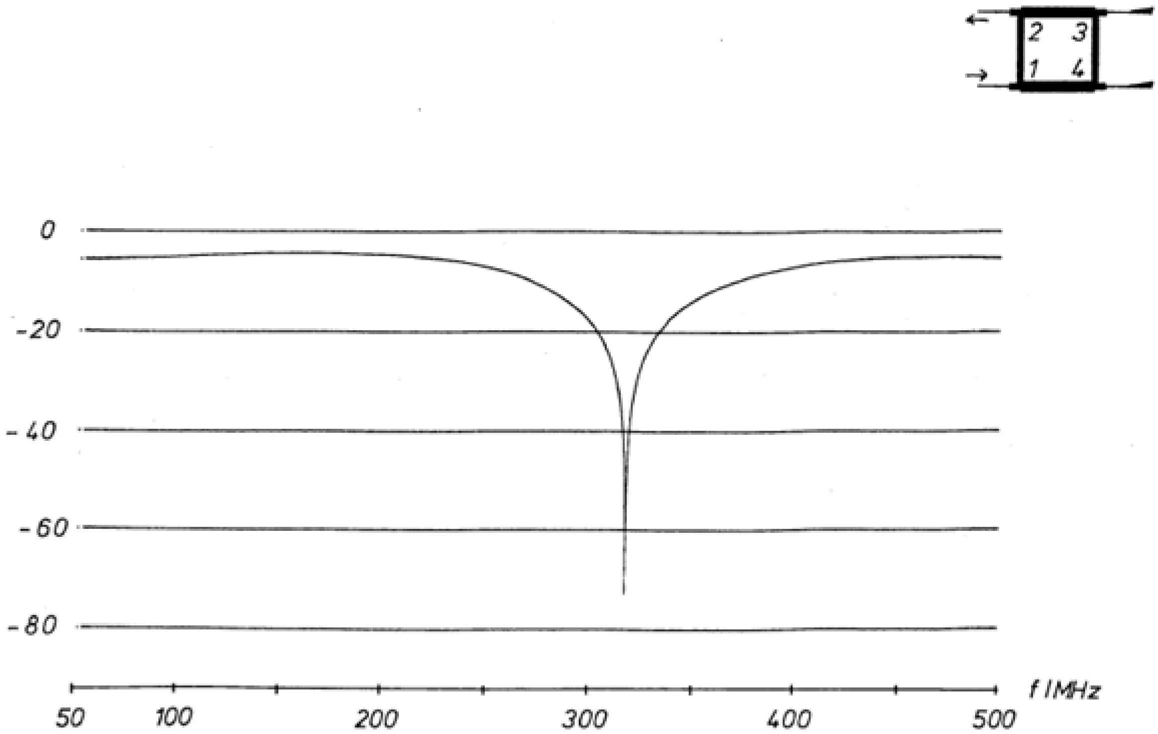


Figure 2.20 Isolation to port 2 reaches a depth of -70 dB, but it is very narrow.

2.4.2 Measurements on the 80m Coupler for 5A7A

During the DX-expeditions to TS7N (2003) and 5A7A (2006) an 80m-coupler design by W2CQH was used.

The table below lists data for this 4-square-coupler (Type D) measured with an R&S vector voltmeter.

	A	B	C	D	E	F	G	H	I	J
1	80 m Koppler Messdaten									
2										
3	f/Mc	South	East	West	Nord	Dummy				
4		In/S dB	In/E dB	In/W dB	In/N dB	In/D dB		Phase S-E	Phase S-N	S-N dB
5	2	-5,9	-0,8	-0,8	-6,3	-25		94	176	0,4
6	2,1	-5,6	-0,9	-0,9	-6	-25		93	175	0,4
7	2,2	-5,3	-1	-1	-5,7	-26		93	175	0,4
8	2,3	-5,1	-1,1	-1,1	-5,4	-26		93	175	0,3
9	2,4	-4,9	-1,2	-1,2	-5,2	-26		92	174	0,3
10	2,5	-4,7	-1,3	-1,3	-5	-26		92	174	0,3
11	2,6	-4,5	-1,4	-1,4	-4,8	-27		91	174	0,3
12	2,7	-4,4	-1,6	-1,6	-4,6	-27		92	174	0,2
13	2,8	-4,2	-1,7	-1,7	-4,4	-27		91	174	0,2
14	2,9	-4,1	-1,9	-1,9	-4,3	-28	3,5 MHz	92	173	0,2
15	3	-4	-2	-2	-4,1	-28		91	173	0,1
16	3,1	-3,8	-2,2	-2,2	-4	-28		92	173	0,2
17	3,2	-3,7	-2,3	-2,3	-3,9	-29		91	173	0,2
18	3,3	-3,7	-2,5	-2,5	-3,7	-29		91	172	0
19	3,4	-3,6	-2,6	-2,6	-3,6	-29		91	172	0
20	3,5	-3,5	-2,8	-2,8	-3,5	-30		90	172	0
21	3,525	-3,5	-2,8	-2,8	-3,5	-30		91	172	0
22	3,55	-3,5	-2,8	-2,8	-3,5	-30		91	171	0
23	3,575	-3,4	-2,9	-2,9	-3,5	-30		91	171	0,1
24	3,6	-3,4	-2,9	-2,9	-3,4	-30		91	172	0
25	3,625	-3,4	-3	-3	-3,4	-31		91	171	0
26	3,65	-3,4	-3	-3	-3,4	-31		91	171	0
27	3,675	-3,4	-3,1	-3,1	-3,4	-31		92	171	0
28	3,7	-3,4	-3,1	-3,1	-3,3	-31		91	171	-0,1
29	3,725	-3,4	-3,1	-3,1	-3,3	-31	Design QRG	91	171	-0,1
30	3,75	-3,3	-3,2	-3,2	-3,3	-31		91	171	0
31	3,75	-3,3	-3,2	-3,2	-3,3	-31		92	170	0
32	3,8	-3,3	-3,2	-3,2	-3,3	-31		91	171	0
33	3,85	-3,3	-3,3	-3,3	-3,2	-32		91	170	-0,1
34	3,9	-3,3	-3,4	-3,4	-3,2	-32		91	170	-0,1
35	3,95	-3,2	-3,5	-3,5	-3,1	-32		90	171	-0,1
36	4	-3,2	-3,5	-3,5	-3,1	-33		91	170	-0,1
37	4,1	-3,2	-3,7	-3,7	-3	-33	3,8 MHz	91	170	-0,2
38	4,2	-3,1	-3,8	-3,8	-2,9	-34		91	170	-0,2
39	4,3	-3,1	-4	-4	-2,9	-34		90	169	-0,2
40	4,4	-3,1	-4,1	-4,1	-2,8	-35		90	169	-0,3
41	4,5	-3	-4,3	-4,3	-2,7	-36		91	168	-0,3
42	4,6	-3	-4,4	-4,4	-2,7	-36		91	168	-0,3
43	4,7	-3	-4,5	-4,5	-2,6	-37		91	168	-0,4
44	4,8	-2,9	-4,7	-4,7	-2,5	-37		90	168	-0,4
45	4,9	-2,9	-4,7	-4,7	-2,5	-38		90	168	-0,4
46	5	-2,9	-4,9	-4,9	-2,4	-38		91	166	-0,5
47	5,1	-2,9	-5	-5	-2,4	-38		90	167	-0,5
48	5,2	-2,8	-5,1	-5,1	-2,3	-38		90	167	-0,5
49	5,3	-2,8	-5,2	-5,2	-2,2	-38		90	166	-0,6
50	5,4	-2,8	-5,4	-5,4	-2,2	-38		90	166	-0,6
51	5,5	-2,8	-5,5	-5,5	-2,1	-37		90	165	-0,7
52	5,6	-2,8	-5,6	-5,6	-2,1	-36		90	164	-0,7
53	5,7	-2,8	-5,7	-5,7	-2	-35		90	164	-0,8
54	5,8	-2,7	-5,8	-5,8	-2	-35		89	525	-0,7
55	5,9	-2,7	-5,9	-5,9	-1,9	-34		90	523	-0,8
56	6	-2,7	-6	-6	-1,8	-33		89	523	-0,9
57	6,2	-2,7	-6,2	-6,2	-1,7	-32		89	522	-1

The design frequency is about 3.8 MHz. A decoupling of -3.2 dB is read on both ports. It is highly dependent on frequency. Decoupling of the isolation port is -31 dB and more than -25 dB between 2 and 6 MHz. The 0.2 dB difference to the theoretical value (-3 dB) is caused by the total losses of the coupler. The phase difference between the output ports stays between 89 and 91 degrees over the entire range. The built-in 180-degree-transformer “rotates” the phase only by 170 degrees in the range of interest.
17 000+ QSOs with 5A7A on 80m were world record in 2006.

3 Power, Voltages, Currents and Losses in a 4-Square System

In my further discussions only the coupler by Reed Fisher will be treated. It has become generally established in the ham world due to its good properties for the "low bands".
The maximum transmittable power of a coupler according to Reed Fischer is determined by its components, such as relays, capacitors, inductors and connectors. Fig. 3.1 shows the circuit diagram of an optimally terminated 3-dB-hybrid-coupler with voltage, current and impedance information.

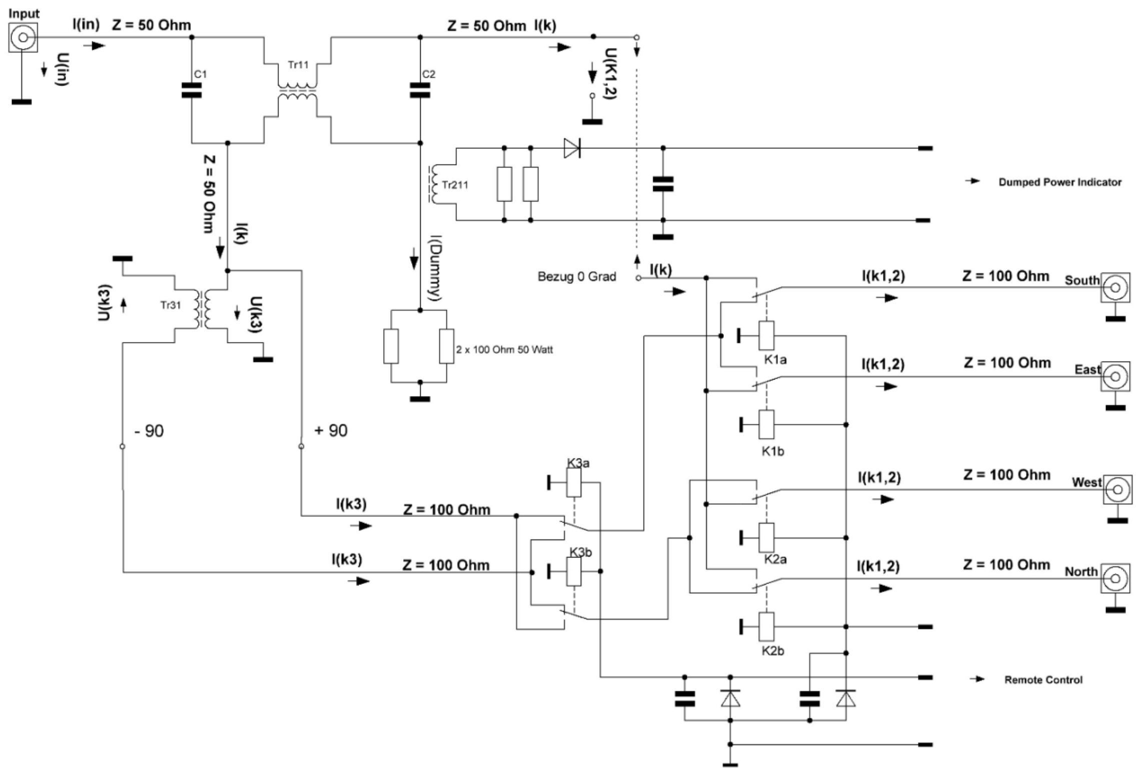


Figure 3.1

3.1. Relays

The table below shows the currents and voltages for different power levels.

Relay Loads (when optimally matched) R(in)=50 Ohm									
	P(in) Watt	U (in) Volt	I (in) Amp	U(K) Volt	I(K) Amp	U (K3) Volt	U(K1,2) Volt	I (K3) Amp	I(K1,2) Amp
	10	22,361	0,447	15,811	0,316	15,811	15,811	0,158	0,158
	100	70,711	1,414	50,000	1,000	50,000	50,000	0,500	0,500
DL	750	193,649	3,873	136,931	2,739	136,931	136,931	1,369	1,369
USA	1500	273,861	5,477	193,649	3,873	193,649	193,649	1,936	1,936
XX	2000	316,228	6,325	223,607	4,472	223,607	223,607	2,236	2,236
XX	5000	500,000	10,000	353,553	7,071	353,553	353,553	3,536	3,536

The relays are selected according to the following criteria:
- power that can be passed through,
- current that can flow through the contacts,
- voltage that the relay can stand

Different to low-frequency applications switching high-frequency power often causes arcing that destroys the contacts. "Hot" switching of high-frequency power should generally be avoided!

Relays with contacts for 220 Volts and 16 Amps are readily available on the market. Since most of these relays have sufficiently good RF-characteristics, they will do for power levels up to 2 KW.

3.2 Capacitors

For capacitors the following parameters are relevant:
- voltage capability and
- loss factor

The max. operating voltage is mostly printed on the capacitors. The loss factor tan δ depends on the dielectric of the capacitor. For high frequency applications suitable dielectrics are air, vacuum, various ceramic materials and mica.
The loss tangent tan δ of good ceramic capacitors for C > = 50 pF ranges between

$$1.5 * 10^{-3} \text{ and } 5 * 10^{-3}$$

for mica capacitors approximately at

$$0.3 * 10^{-3}$$

$tan\delta$ is the ratio of real and reactive power

$$Pw = \frac{1}{2} * |U|^2 * Gp$$

$$Pb = \frac{1}{2} * |U|^2 * \omega C$$

$$tan\delta = \frac{Pw}{Pb} = \frac{Gp}{\omega C}$$

Real power in a capacitor is converted into heat. As mica capacitors dissipate less than 0.3 W real power at 2 kW, they are the best choice. Because of the power splitting1 kW of the total power are transported to the second port via the capacitor. For this power level capacitors for 500 volts are sufficient.

3.3 Inductors

In the couplers iron powder cores with mix 2 (red) material are used. They have a permeability of 10. Compared to other iron powder cores the relative costs are the lowest. Cores with material mix 6 (yellow) have a permeability of 8.5, are thermally more stable and have lower losses, but are more expensive. They all consist of very fine carbonyl iron powder embedded in a carrier material with distributed air gaps. These cores have sufficient permeability, are temperature stable and can be used to build inductors and transformers of high quality. In applications with sinusoidal signals, the max. transferable power is not determined by possible saturation, but by heating due to winding and core losses.

3.3.1 Skin Effect

The winding losses depend on the RF-resistance of the wire. With increasing frequency the effective conductor cross-section is reduced by the influence of the skin effect:
Skin Depth (mm) = 66.2 / √ f
The skin effect has less influence on LF. Nevertheless, for low losses the largest possible conductor cross-section should be chosen for inductors.

3.3.2 Core Losses

Core losses depend on the core material. For material 2 they are illustrated in the diagram of Micrometals below. Losses can be calculated with the formula given in the diagram. The formula for magnetic flux density B runs:

$$B = \frac{E * 10^8}{4.44 * A * N * f}$$

with

"Power Rating" for 25° C temp rise due to core loss	
Material 2 – 1 MHz	
Core	Watts
T30	21
T37	26
T44	37
T50	49
T68	88
T80	125
T94	160
T106	236
T130	331
T157	515
T200	794
T300	1127
T400	2108

FIGURE K

B = magnetic flux density in Gauss
E = effective voltage in Volts
A = core cross section in cm²
N = number of turns
f = frequency in Hertz

With the formulas in the applications notes of Micrometals core losses can be calculated for various frequencies, voltages, cores and windings.
For example, with an assumed power of 1000 watts the voltage is 224 volts at 50 ohms. For an estimate I take the voltage across a winding of a core of type T200-2 with a core cross-section of 1.333 cm² at a frequency of 3.65 MHz and 14 turns. With these values the flux density B amounts to 75 Gauss, very far from the max. saturation flux density of 3000 Gauss of this material. Substituting the obtained values in the formula for the core losses (cf. diagram) in a core volume of 17.25 cm^3 core loss is 5.8 watts

.

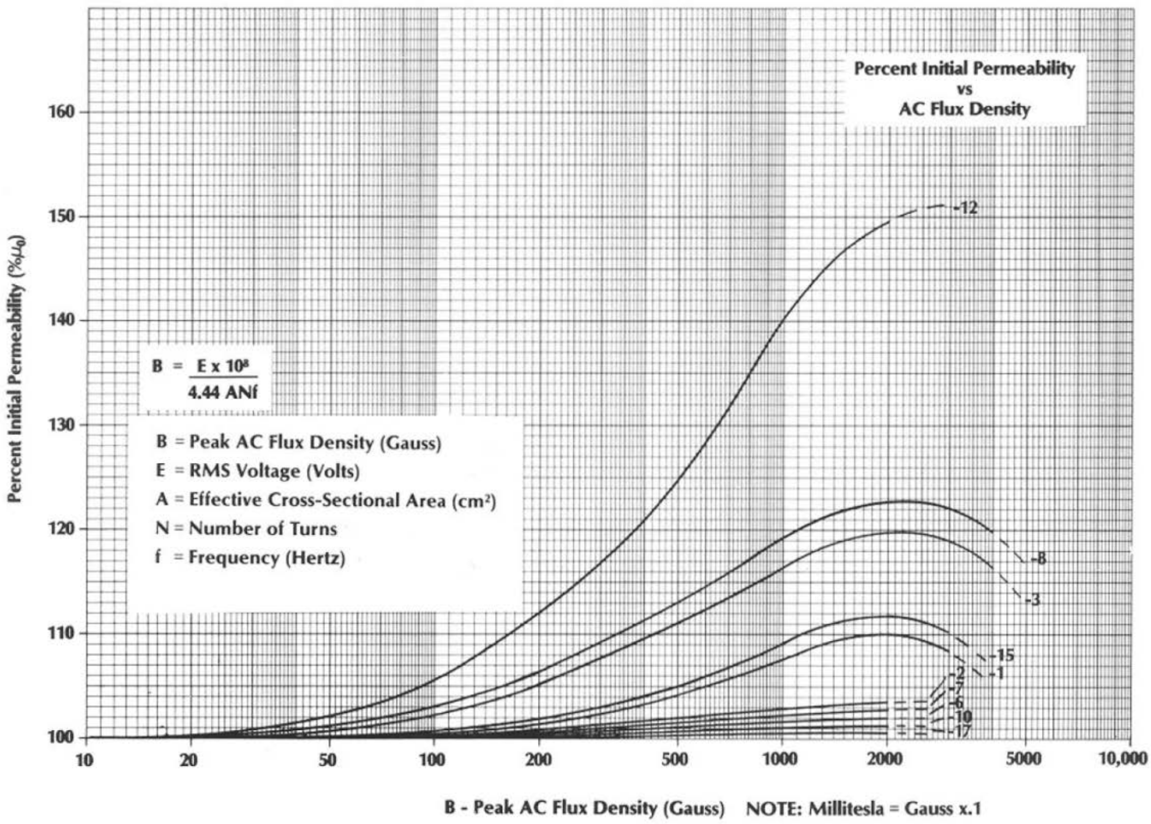


Figure 3.2

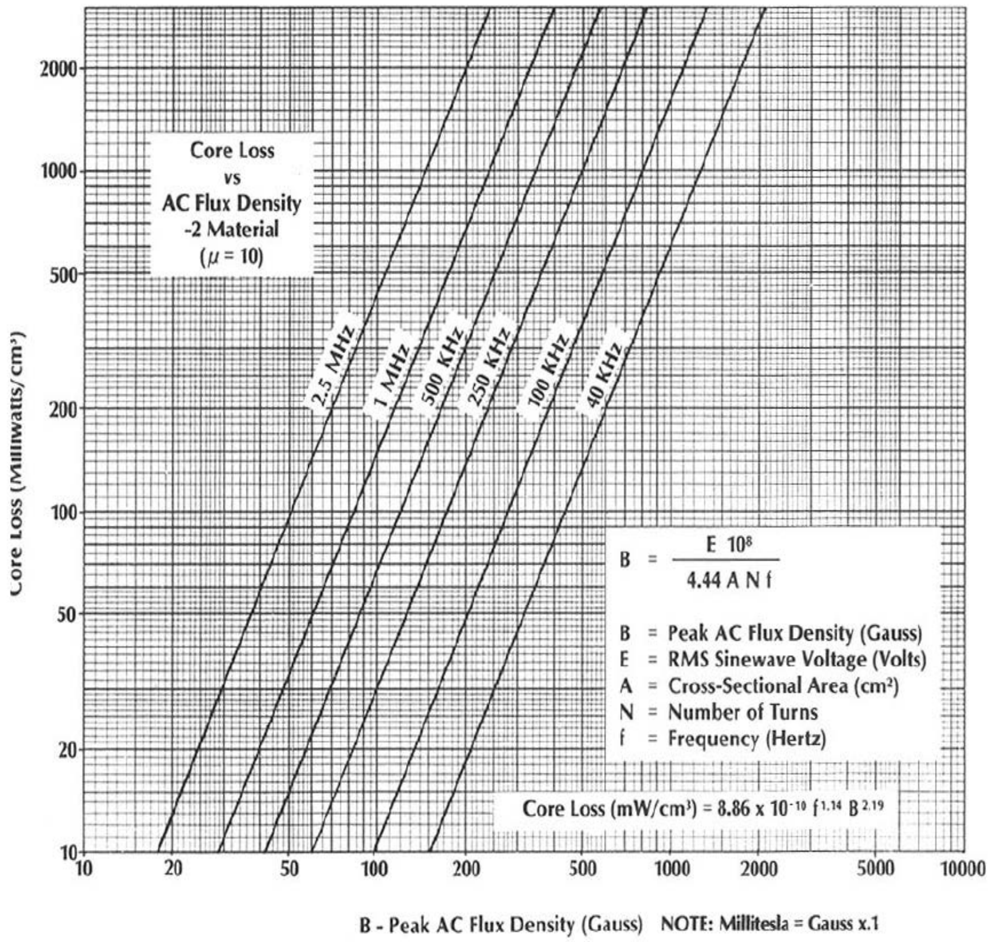


Figure 3.3

The diagrams in Figures 3.2 and 3.3 are taken from Micrometals' application notes.

The table above shows the power that will cause a temperature increase of 25 degrees in material 2 cores at a frequency of 1 MHz. It is the max. permissible continuous power for the different cores listed in the table.

For the selection of cores saturation does not matter. The decisive factor is the max. allowed core temperature mainly by loss in the core and the winding. For the selection of the core the average power during operation time must be considered after core, winding, housing, ventilation, etc. have been thermally stabilized. As the average power in CW is about twice as high as in SSB, you can specify double the max. CW-power for the 4-square-coupler in SSB. For a rough estimate of a temperature increase of 25 degrees due to core losses we choose two T200-2 cores for the hybrid. According to the table about 1600 watts of continuous power can be used at 1 MHz. With a duty cycle of 1:1 about 3200 watts would be allowed for CW and about 6400 watts for SSB. With the two T200-2 cores you are on the safe side for amateur radio applications. In the applications notes of Micrometals Jim Cox has published two most interesting articles:
"Iron Powder Core Selection For RF Power Applications" and
"Iron Powder Cores for High Q Inductors"

Here you will find a solid basis for the selection of iron powder cores for high frequency applications. If voltages, frequency, number of turns and core type are known, core losses can be easily calculated to estimate the resulting temperature increase.

3.4 Radiator Radial System

The feed-point impedance of an antenna system is determined by the sum of the radiation resistance and all losses:

$$Z_{in} = Z_s + Z_l$$

Losses must be kept as low as possible. Ohmic losses in the radiator are determined by its dimensions, the conductivity of its material and the skin effect at the operation frequency. If ground planes are selected as radiators, losses are essentially earth losses. The latter are largely determined by the radial system. This has been very well studied by N6LF and OE3REB in their contributions: http://oe9.oevsv.at/export/oevsv/download/QSPArchiv/2009/QSP200910.pdf
http://www.antennasbyn6lf.com/2009/12/series-of-qex-articles-on-ground-system-experiments.html

3.5 Cable Losses

On the low bands cable losses are quite low. They are determined by the length and type of cable and the SWR on it. Cable data and the influence of SWR can be found in most antenna books.

3.6 Sum of Losses

Total loss consists mainly of the following partial losses:

losses of the coupler
- the $tan\delta$ of the capacitors
- resistive and skin effect losses of windings and PCB strip lines
- core losses of the inductors
- contact resistance of contacts, connectors, relays
- power into the dummy load at mismatch

cable losses in λ/4-transformers
radiator and radials
- ground losses
- matching losses
- radiator material losses

Try carefully to minimize losses.
The laws of physics cannot be outwitted - all losses sum up unmercifully!

4. Reed Fisher (W2CQH) Coupler Design

4-square antennas require a power split of -3 dB to two ports with a phase difference of 90 degrees. W2CQH has described the technique in QST, Jan. 1976, in his article "Twisted-Wire Quadrature Hybrid Directional Coupler". This type of coupler has become popular because it is simple and doesn't require much space on the low bands.

4.1 System Requirements

1. 50 percent of the power from the input port 1 to port 2
2. 50 percent of the power from the input port 1 to port 4
3. 0 percent of the power from the input port 1 to port 3
4. no reflections at the input port 1; input SWR = 1
5. the phase angle of voltage at port 2 and port 4 differs by 90 degrees

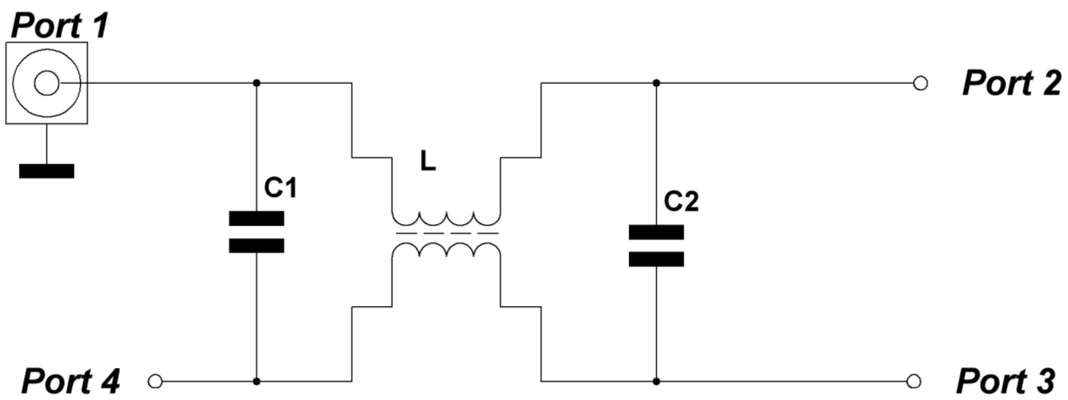


Figure 4.1

In theory conditions 3, 4, and 5 are independent of frequency, but conditions 1 and 2 are strongly frequency dependent. Near the design frequency the coupler works as a matched power splitter with a 90 degrees phase difference of the output voltages.

4.2 Calculation Formulas

- The characteristic impedance of the system is:

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

- For an equal splitting of power applies:

$$Z_0 = 2\pi f_0 L$$

$$Z_0 = \frac{1}{2\pi f_0 C}$$

resulting in:

$$L = \frac{Z_0}{2\pi f_0}$$

$$C = \frac{1}{2\pi f_0 Z_0}$$

C is the total capacity of the system. Specific capacities are calculated by:

$$C_1 = C_2 = \frac{C - C_x}{2}$$

C_x = stray capacitances

For the inductivity of a toroid applies:

$$\frac{L}{N^2} = \frac{4\pi\mu A}{l}$$

L = inductivity (nH)
µ= permeability
A= core cross-section (cm²)
l = length of the field line (cm)
N = number of turns

Core permeability, core cross-section and field line length are summarized by the manufacturers as the AL-value. Use AL [nH / turns^2] and [nH] for L as units in the calculation formula.
Calculate the number of turns from:

$$N = \sqrt{\frac{L}{A_L}}$$

4.3 Useful Dimensions

For160, 80, 40 and 30m the values for capacities and inductivities are listed below.

Capacity and inductivity

	C = C1=C2	L
f/MHz	C/pF	L/nH
1,85	861	4304
3,65	436	2181
7,05	226	1129
10,10	158	788

Mica capacitors are available with very tight tolerances. In order to tune the coupler, splitting the total capacity into several capacitors is advisable. By connecting capacitors in series the voltage capability of the total capacitance is increased.

Number of turns for some iron powder cores

	Core parameters			frequency in MHz			
Core type	AL/nH/N²	A/cm²	V/cm³	1,85	3,65	7,05	10,1
				number of turns			
T157-2	14	1,06	10,7	17,5	12,5	8,9	7,5
2*T157-2	28	2,12	21,4	12,4	8,9	6,3	5,3
T157-6	11,5	1,06	10,7	19,3	13,8	9,8	8,3
2*T157-6	23	2,12	21,4	13,7	9,8	6,9	5,9
T184-2	24	1,88	21	13,4	9,6	6,8	5,7
T184-6	19,5	1,88	21	14,8	10,6	7,5	6,4
T200-2	12	1,27	16,4	18,9	13,5	9,6	8,1
2* T200-2	24	2,54	32,8	13,4	9,6	6,8	5,7
T200-6	10,4	1,27	16,4	20,3	14,5	10,3	8,7
2* T200-6	20,8	2,54	32,8	14,4	10,3	7,3	6,2
T200-2B	21,8	2,32	30	14,0	10,0	7,1	6,0
T225-2	12	1,42	20,7	18,9	13,5	9,6	8,1
2*T225-2	24	2,84	41,4	13,4	9,6	6,8	5,7
T225-6	10,4	1,42	20,7	20,3	14,5	10,3	8,7
2*T225-6	20,8	2,84	41,4	14,4	10,3	7,3	6,2
T225-2B	21,5	2,59	37,8	14,1	10,1	7,2	6,1

Core material 2 is slightly cheaper than material 6, but it has slightly higher core losses. The permeability of material 2 is 10, that of material 6 is 8.5.
The AL-value of cores is generally specified with a tolerance of ± 5%. See:
http://www.micrometals.com/materials_index.html
To determine the number of turns the “Mini Ring Core Calculator” of DL5SWB can be used. Its installation file can be found at
http://www.dl5swb.de/html/mini_ringkern-rechner.htm.
As can be seen in the table above, integer winding numbers are rare. The desired inductance can be influenced only limited by adjusting the distribution of the turns on the toroid.
Refer to:
http://www.micrometals.com/appnotes_index.html
under RF Applications by Jim Cox
1) “Iron Powder Cores for High Q Inductors”
by Jim Cox - Micro Metals, Inc. , page 8
2) “Iron Powder Core Selection for RF Power Applications”
by Jim Cox, Micro Metals, Inc.
In these two PDF-files Jim Cox has competently put together very interesting information for the applications of iron powder cores.
According to my practical experience the inductance essentially determines the input impedance of the coupler. Since you should choose large cross-sections of the

winding wire to reduce skin effect losses, adjustment of turns on the core to trim the inductance is difficult.
If you use the capacity values from the table you may see that the signal amplitudes at the output ports are not equal. This can be corrected by slightly varying the capacitance values to obtain the same amplitudes at the ports. The capacity values may slightly differ from the calculated values. Therefore, it is useful to provide for adjustments on the PCB (Printed Circuit Board). To familiarize yourself with the various factors, it is recommended to build a sample of the coupler with values from the table for the design-frequency and to terminate the ports with 50 Ohms. Now the fine adjustment of the output amplitudes can easily be made.

5 System Impedance - Bandwidth - SWR - Dumped Power

Fig. 5.1 shows a 4-square antenna system consisting of radiators with radial systems, quarter-wave transformers, current-baluns and the phase-shifting and power-splitting network. The different coupler types have a different behavior regarding power splitting and phase shifting. The coupler according to Reed Fisher with a constant phase shift of 90 degrees and an amplitude error of about 0.2 dB at the 80m-band edges appears to be the best choice. All further discussions refer to this type of coupler.

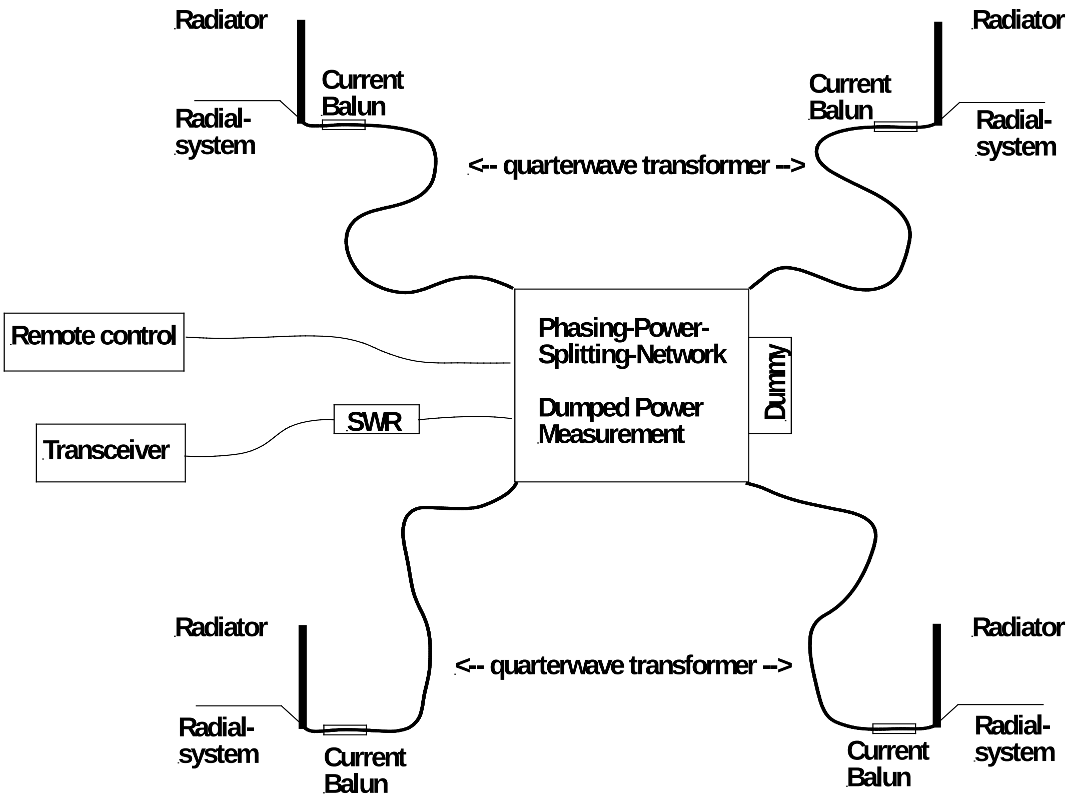


Figure 5.1 Schematic diagram of a 4-Square Array

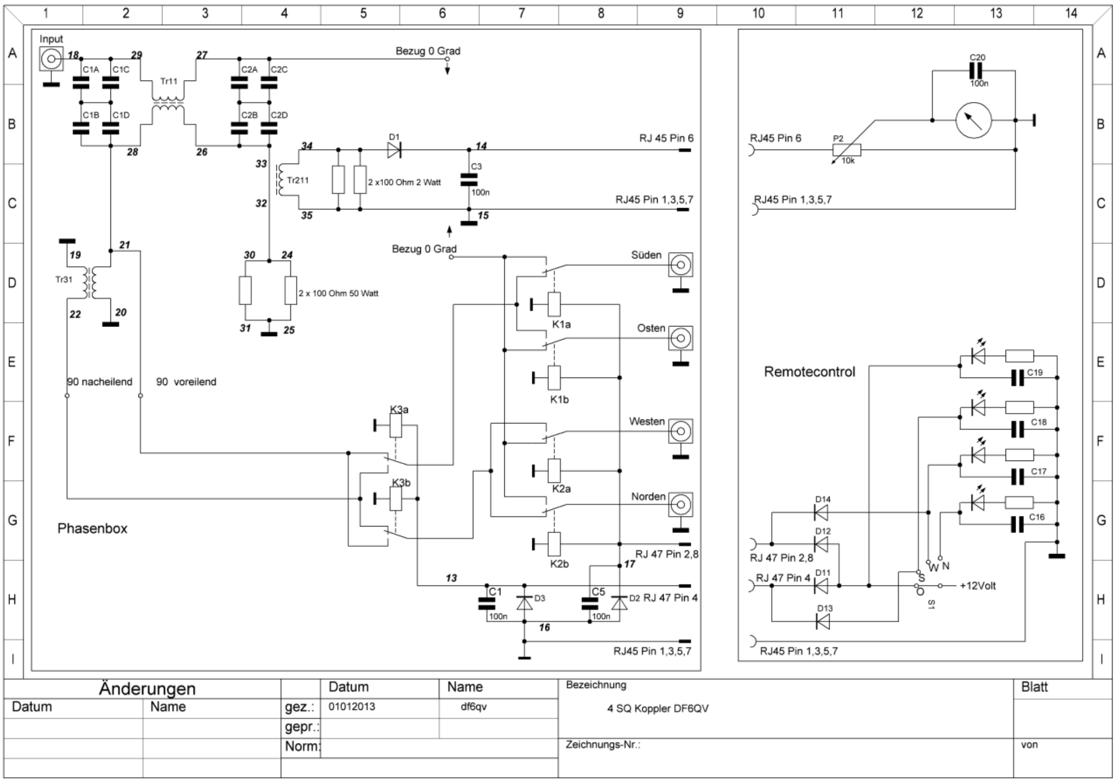


Figure 5.2 Circuit diagram of the coupler and control panel

5.1 System Impedance

Maximum energy transfer occurs with the optimal match of all individual system components. A system impedance of Z = 50 ohms ± j0 ohms has been chosen, as it is the common impedance of the transceiver output. All of the following components should have an impedance as close as possible to this system impedance.
First of all, the phase-shifting and power-splitting network is designed for this impedance. As there will be rarely an integer number of turns from the calculation of the inductances, you will hardly meet the system impedance accurately. In my experience, the inductance essentially determines the input impedance of the 3 dB hybrid, e.g. for a 7 MHz hybrid with two type T200-2 cores 6.8 turns are required. According to manufacturers' specifications the cores have a tolerance of about ± 5%. Selecting 6 turns results in an input impedance of about 46 ohms, with 7 turns one of 55 ohms, both causing an input SWR of about 1.1. The 3-dB-hybrid is adjusted with the capacity at the center frequency. At the output ports of the 3-dB-hybrid there are two signals differing by 90 degrees, each with -3 dB. Losses are disregarded here. Two radiators each are connected at the ports, two directly in parallel, the other two via a 180-degree-transformer. For a good match the four outputs of the system must be terminated with 100 Ohms. The radiators are connected to them via 75Ω-λ/4-lines transforming the 50 Ohm-feed-impedance of the radiators to the required 100 Ohms. If the feed-impedance of the radiator is different from 50 Ohms, it must be transformed to 50 ± j0 Ohms for an optimal match. The λ/4-lines should be made

from coaxial cable with a velocity factor ≥ 0.7, otherwise a radiator distance of λ/4 cannot be obtained. For closer element distances see Sec. 1.3.3. The 80m-4-square at 5A7A, for example, was operated with PE (VF = 0.66) isolated λ/4-lines.
All reflected power from the radiators is dissipated in the dummy load of the 3-dB-hybrid. This power is lost for transmission. Equally the reception power of the 4-square is reduced as well. In the 4-square at 5A7A with a resonant frequency at 3.4 MHz about 10% of the TX output power were wasted in the dummy load. The signal could have been about 1 dB louder. The antenna could not be properly adjusted because the third harmonic (3750KHz) of an nearby AM station on 1250 kHz prevented any antenna measurements. It should be noted that despite the imperfectly tuned radiators the SWR into the coupler was nearly 1.0. The resonance frequency of the system could be seen from the minimum reflux into the dummy load at 3.4 MHz. The SWR into the -3db-hybrid-coupler is not a measure of the quality of the antenna! The quality of the antenna system can only be assessed from a low reflux into the dummy load of the coupler.
It is a "must" to measure the power that is directed to the dummy load!

5.2 Bandwidth

The bandwidth of an antenna system is the frequency range in which the electrical parameters change within pre-set values. The following parameters are to be considered for a 4-square array:
- input SWR
- frequency-dependent power split of the coupler
- frequency-dependent phase difference
- ratio of the input power vs. the power absorbed in the dummy load
- frequency-dependent transformation of λ/4-lines
- frequency dependence of the radiating elements

Summarizing "bandwidth" in a single term will hardly be possible. It makes more sense to optimize the system for the operating frequency and to adapt it by frequency-dependent switching of the antenna elements the system.

5.3 SWR

The input impedance of the coupler is determined by the dimensions of its elements. The input SWR at the system impedance, generally 50 Ohms, can be calculated or measured. This is true for ideal -3dB-couplers, if the output ports are terminated with equal impedances (not necessarily 50 Ohms). For these conditions the reflected power at the output ports is totally directed to the dummy load.

The input SWR is not a measure of the quality of the antenna system!
The SWR is even close to 1.0, if all output ports are open or shorted!
It is important to minimize the dumped power by matching the antenna elements to the output ports of the coupler!

5.4 Isolation - Power Splitting - Phase Difference

In the simulation (cf. Fig. 2.9) the isolation of the input to the dummy load port is better than 40 dB over a wide frequency range. In the real world, the insulation is frequency dependent. For the 160m-band I have measured values of > 30 dB, for the 80m-band > 27 dB and > 20 dB for 40m. Frequency-dependent losses in the overall design have a decisive influence. The phase difference between the output ports is 90 degrees over a wide range. For the generation of the 180-degree phase shift a 180-degree transformer is used. In practice a phase shift of 180 degrees cannot be fully achieved. The maximally possible phase difference depends on the cores, the number of turns and the winding technique. According to my experiences transformers are useful up to a frequency of 7 MHz. For higher frequencies a λ/2-line is the better choice to generate a 180-degree phase difference.
Different to the dimensions of the radiating elements the coupler has only little influence on the bandwidth of the overall system.

5.5 Dumped Power

Power reflected from the radiators is absorbed in the dummy load. If the radiators are precisely matched to the system impedance of the coupler, the reflected power approaches to zero. Antennas for our amateur bands have a real impedance (often different from the system impedance) only in a small frequency range. For the 80m band with a relative bandwidth of 8% in Region 1 the band should be divided into three segments in order to adapt the radiators optimally to the system. In the center of each segment the radiators or the radials should be adjusted to an impedance close to $Z = 50\ \Omega \pm j0\ \Omega$ by changes in length or by suitable matching circuits.

5.6 Current-Balun

A current balun can be made by running coaxial cable through a number of suitable ferrite beads that form an inductivity on the outside of the shield of the coax. The AC-resistance of the inductivity “chokes” the common mode current on the outer side of the shield of the coaxial cable. The lower the common mode current, the less radiation from the feed-line. “Common-Mode Transmission Line Currents” are described in "The ARRL Antenna Book" 19th Edition, pp. 26-16. Radiation from the feed-line is superimposed on the radiation of the antenna. It should be small to avoid distortion of the calculated radiation pattern. The AC-impedance of the balun should be at least 300 Ohms for an impedance of 50 Ohms at the feed-point. The current balun should be installed close to the feed point of the antenna, since the ratio of inductive reactance to the common mode impedance is optimal there.

5.7 Mutual Coupling

The electromagnetic interaction between antenna elements in an antenna array is referred to as mutual coupling. This applies to spacings of the antenna elements within the near field. More details can be found in the PDF-file of Hon Tat Hui at http://www.ece.nus.edu.sg/stfpage/elehht/Receiving%20Mutual%20Impedance/Mutual%20Coupling%20in%20Antenna%20Arrays.pdf
For the “normal” radio amateur (even for me) the paper is very mathematical and complex. I prefer a simulation of a 2-element array with an element spacing of λ/4

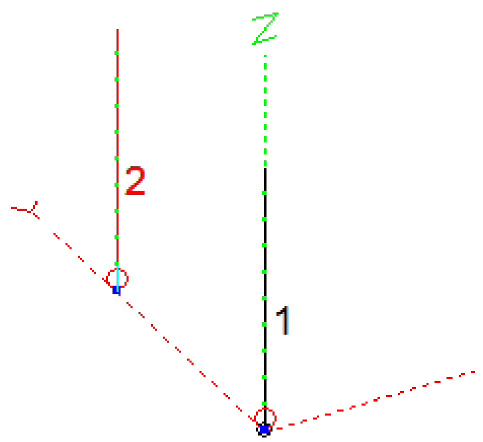
that will show, how the feed impedance is affected by mutual coupling. Figs. 5.3 - 5.6 show the array data in EZNEC.

Wires										
No.	End 1				End 2				Diameter	Segs
	X (m)	Y (m)	Z (m)	Conn	X (m)	Y (m)	Z (m)	Conn	(mm)	
1	0	0	0	Ground	0	0	10,3		40	10
2	0	10	0	Ground	0	10	10,3		40	10

Figure 5.3

Sources							
No.	Specified Pos.		Actual Pos.		Amplitude	Phase	Type
	Wire #	% From E1	% From E1	Seg	(V, A)	(deg.)	
1	1	0	5	1	1	0	I
2	2	0	5	1	0	90	I

Figure 5.4



File	Vert1.ez
Frequency	7 MHz
Wavelength	42,8275 m
Wires	2 Wires, 20 segments
Sources	2 Sources
Loads	0 Loads
Trans Lines	0 Transmission Lines
Transformers	0 Transformers
L Networks	0 L Networks
Ground Type	Real/MININEC
Ground Descrip	1 Medium (0,005, 13)
Wire Loss	Zero
Units	Meters
Plot Type	Elevation
Elevation Angle	0 Deg.
Step Size	1 Deg.
Ref Level	0 dBi
Alt SWR Z0	75 ohms
Desc Options	

Figure 5.5

Figure 5.6

The impedance graph of element 1 with 1 amp feed-current is shown in Fig. 5.7. . With Element 2 unfed its impedance is Z = 37.5 + j10 ohms at 7.1 MHz.

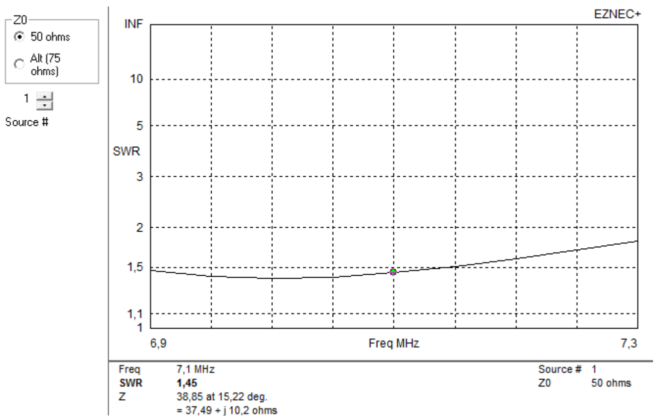


Figure 5.7

The graphs in Figs. 5.8 and 5.9 show the feed-point impedances of both elements when Element 2 is also fed with1 amp but with a phase-shift of 90 degrees. Element 1 has now an impedance of 53+j38 and Element 2 one of 22- j11.5.

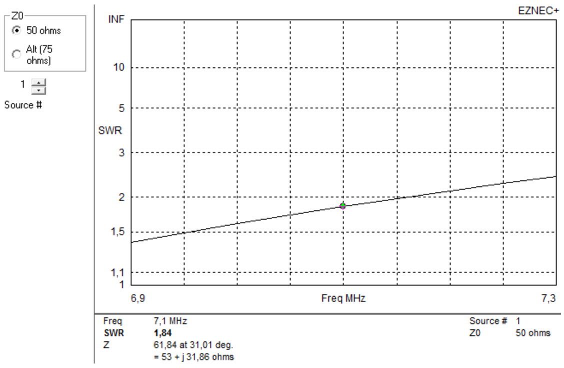


Figure 5.8

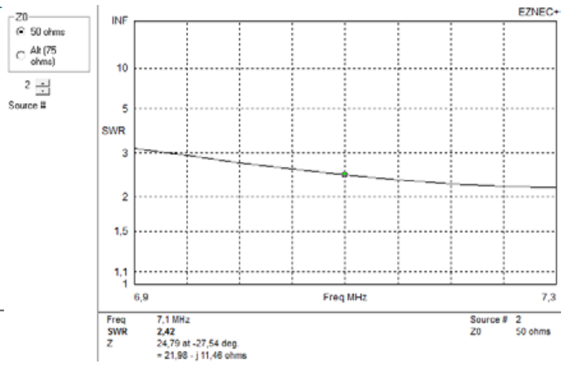


Figure 5.9

It can be seen that the coupling of the elements has an influence on the impedance of the two elements.

5.8 Power Match

Optimal match is achieved when the output impedance is equal to the input impedance of the following system. But since in most cases the imaginary part of the impedances is not equal to zero the output impedance must be complex conjugate to the input impedance of the following system.

Re (Zout) = Re (Zin) and Im (Zout) = - Im (Zin)

For matching the transmitter to the coupler impedance can be assumed as almost real. The coupler input has a very flat SWR over a wide frequency range. The circuit diagram in Fig. 3.1 shows the impedances inside the coupler. The four outputs have an impedance of 100 Ohms, as two phasing-lines are connected in parallel forming a 50 Ohm load for the coupler ports. Therefore, the 50 ohm feed-point impedance at the four radiators must be transformed to 100 ohms with 75-ohm-λ/4-lines. The radiators have a real input impedance of 50 ohms only in a small frequency range and due to mutual coupling the impedances of the antenna affect each other (cf. section 5.7).

5.9 Current Forcing

The principle of current forcing is described in "The ARRL Antenna Book" 19th Edition, pp. 8 -15. In λ/4-lines the output current is equal to the input voltage divided by the impedance of the λ/4-line, regardless of the magnitude of the load impedance. Output current and input voltage have a phase difference of 90 degrees. That is why equal currents are forced in elements with different impedances, eg. caused by different ground characteristics.
A measure for the energy flux density in the ether is the Poynting vector S with the

unit Watts/area. It is the cross product of the electric and magnetic field strength. The electric field strength E and the magnetic field strength H are functions of the voltage and current that are linked by the radiation resistance of the antenna.

5.10 λ/4-Line-Transformers

To transform the 50 Ohm feed-point impedance of the radiators to 100 Ohms at the ports of the phasing-box we choose λ/4-lines with a characteristic impedance of 75 Ohms. Two lines each are connected to the ports in parallel resulting in totals of 50 Ohms at the output ports of the coupler. To calculate the required characteristic impedance of the lines we use the formula:

$$Z_{line} = \sqrt{Z_{radiator} * Z_{port}}$$

With $Z_{radiator}$ = 50 Ohms and Z_{port} = 100 Ohms

$$71 = \sqrt{50 * 100}$$

Z_{line} should ideally be 71 Ohms, pretty close to 75 Ohms.

Since you cannot obtain λ/4-spacing of the radiators with a PE-insulated coaxial cable, foam-insulated cable is preferable. For smaller spacing cf. Chap. 1.3.3.
The λ/4-lines serve two purposes: they force the output current that is equal to the input voltage divided by the impedance of the line and they transform the feed-point impedance according to the formula above.

6 4-Squares in Practice

6.1 Radiators

Ground planes with ground radials or one or two elevated radials are generally used as elements for 4-squares. If no high-angle radiation is desired, folded elevated radials (FCP) can be used. Cf. http://www.w0uce.net/FCP_Facts.pdf
or http://www.oevsv.at/export/oevsv/download/QSPArchiv/2009/QSP200910.pdf
(page 53 ff.).
Each radiator should have an impedance of Z = 50 ± j 0 Ohms. Above ideal ground the base impedance is 36 Ohms. If this is the case, the antenna should be matched to 50 Ohms for optimum performance. Suitable matching circuits, e.g. Smith V3.10, can be downloaded at http://www.fritz.dellsperger.net/ .
A vertical λ/4-radiator with one elevated radial 1 to 1.5 meters above ground has a feed-impedance close to 50 ohms at 3.5 MHz. The impedance is largely determined by the geometry of the radiator and by ground losses. It can be considered as a bent dipole with a vertical and a horizontal leg. By changing the length of the horizontal leg you can adjust the resonance of the antenna.

Fig. 6.1 shows the 40m 4-square at DF0AT using four 12m-Spiderbeam poles with one elevated-radial for each radiator.

Figure 6.1 40m 4-square with one elevated-radial at DF0AT

6.2 Current Balun

In Fig. 6.2 the feed-point of a radiator of a 40m 4-square with a current balun made of 8 ferrite cores can be seen. In Fig. 6.3 the foot of an 80m-vertical is shown. To choke common mode currents two wires (purple and black) are guided through four ferrite cores several times. The red coil (Z_L > 500 ohms) shorts static charges against ground. The common mode choke should have an impedance of > 300 ohms on the operating frequency and must be positioned close to the base of the radiator.

Figure 6.2

Figure 6.3

Figure 6.4 SWR of a 40m 4-square element on 7064 ±200 KHz

Figs. 6.5 - 6.7 show details of the measured data of a 30 core common mode choke. Common mode currents on the outer conductor should be as small as possible. The arrangement with 30 ferrite beads has an attenuation of 14 dB at 1.8 MHz, of 16.2 dB at 3.5 MHz and of 17 dB at 7 MHz (cf. Fig. 6.7). Unknown cores should first be tested for their performance (see attenuation measurements Figs. 6.5 - 6.7).

Figure 6.5

Figure 6.6

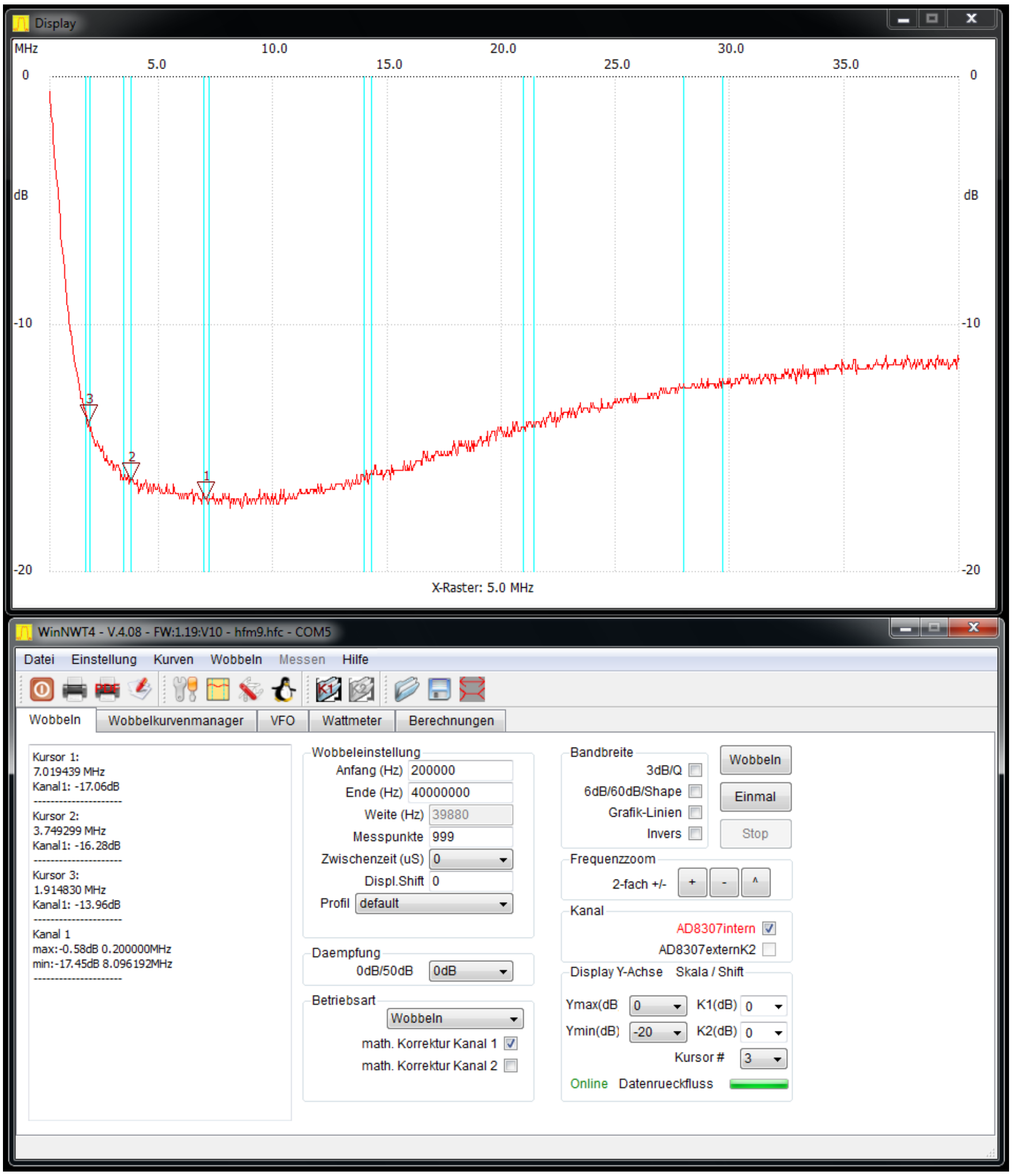


Figure 6.7

The screenshot of the measurement of a current balun made of 8 ferrite cores from PC-monitor cables can be seen in Figure 6.8.

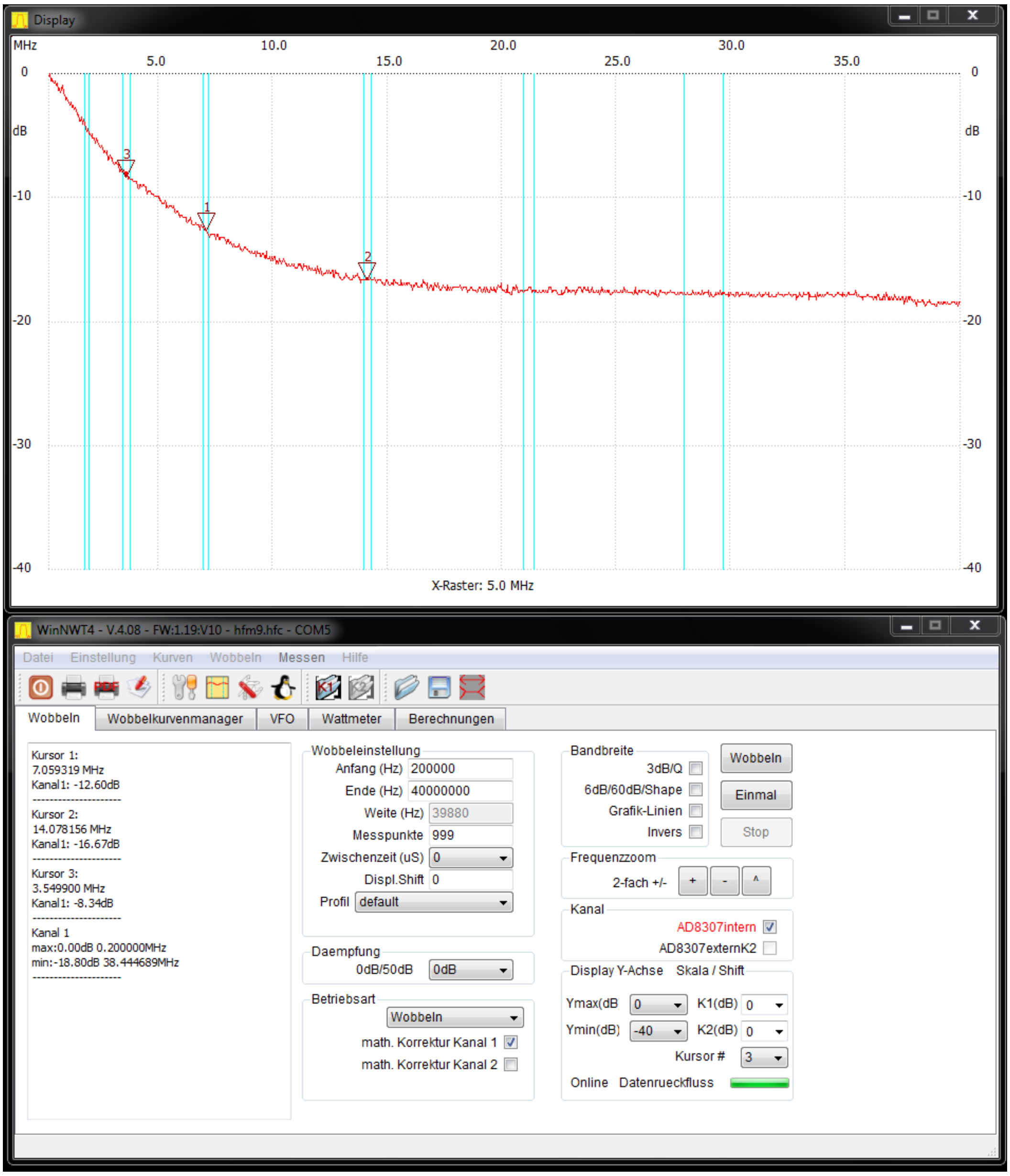


Figure 6.8

Figure 6.10 shows the effect of ten turns of RG 213 with a diameter of about 15 cm on current mode currents.

Figure 6.9 Ten turns of RG213 as a current balun

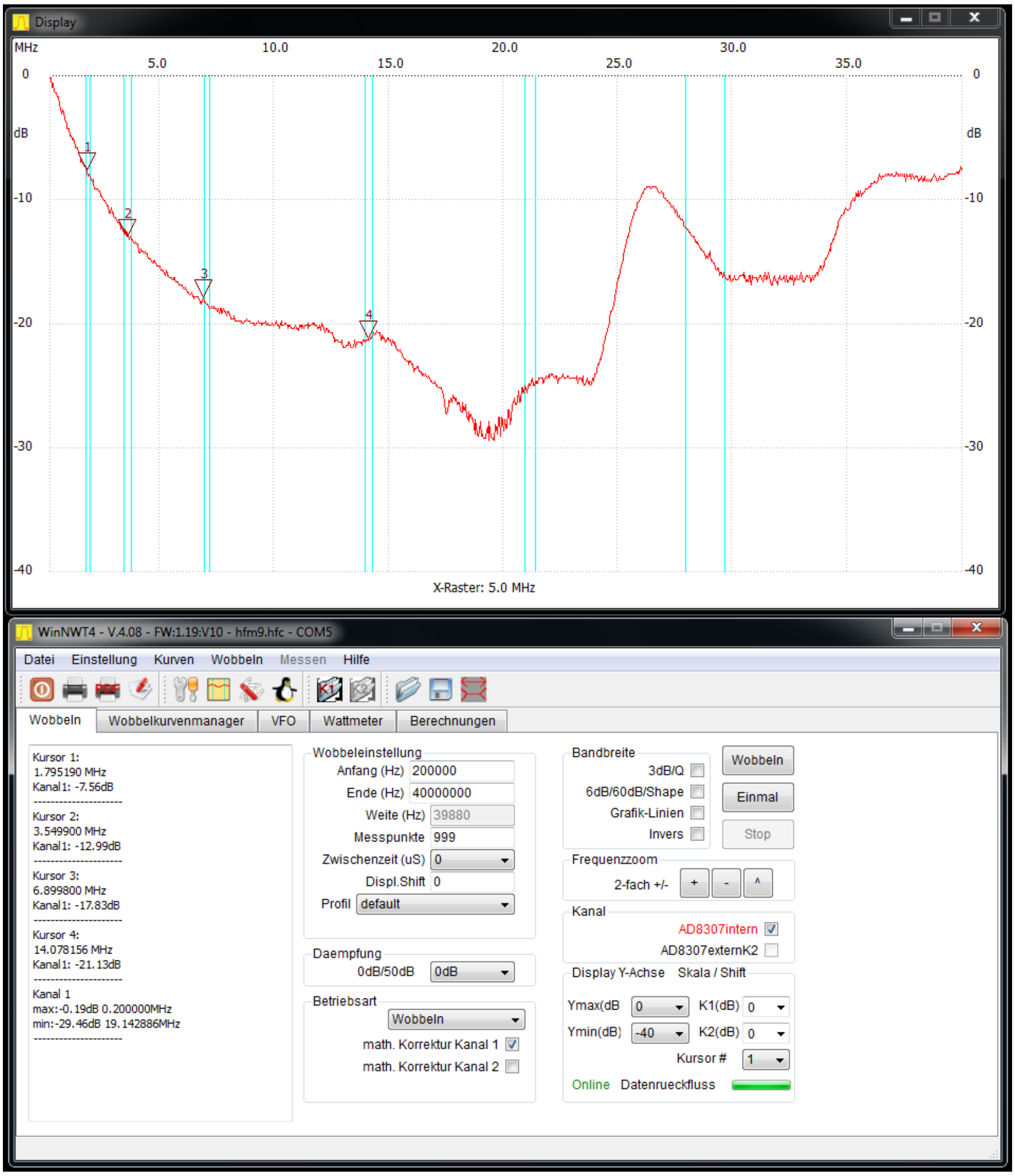


Figure 6.10

6.3 The Phasing Box

The phasing-box splits the power of the transmitter into four equal parts with the required phases. It is designed for a 50 Ohm system. Of the input power twice 25 per cent can be decoupled from each of the two output ports. Less than -20 dB of the input power go into the dummy load, if the output ports are terminated properly.

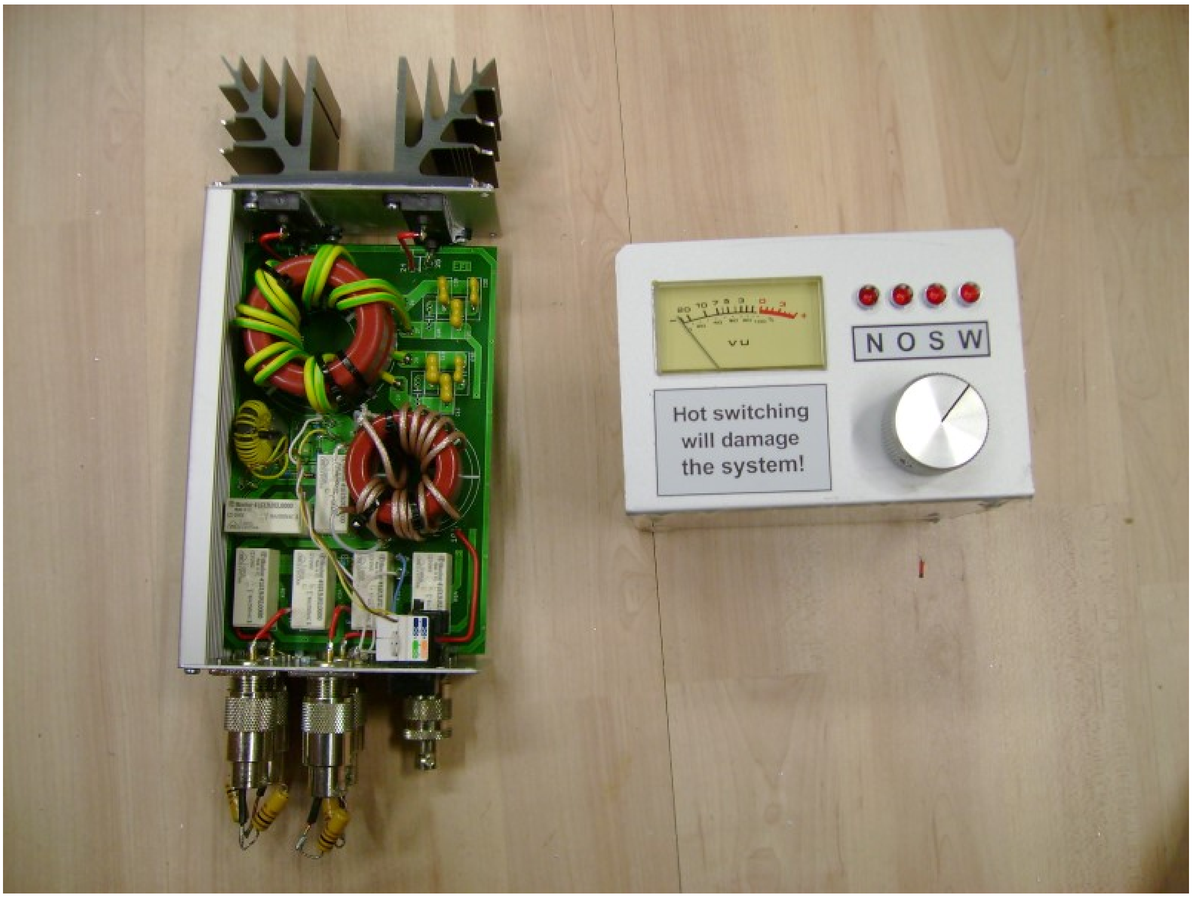


Figure 6.11

In Fig. 6.11 the interior of a coupler is shown on the left. The output terminals are terminated with 100 Ohm resistors for the purpose of measurement. Fig. 6.12 shows the input SWR. The decoupling of the dummy load and the phase relationship of the output signals are to be seen in Figs. 6.14 and 6.15. On the right in Fig.6.11 you can see the remote control box with the power meter of the dummy load. For 100 Watts into the dummy load the meter is adjusted to 0 dB, for 50 Watts -3 dB, 25 Watts -6 dB etc. to be read on the display that is scaled in dB.
The input SWR of the phasing-box is very flat over a wide range, since it is mainly determined by the dimensions of the coupler. Even with open or shorted outputs the SWR remains small.
The input SWR of the phasing-box is NOT a measure of the quality of the antenna!

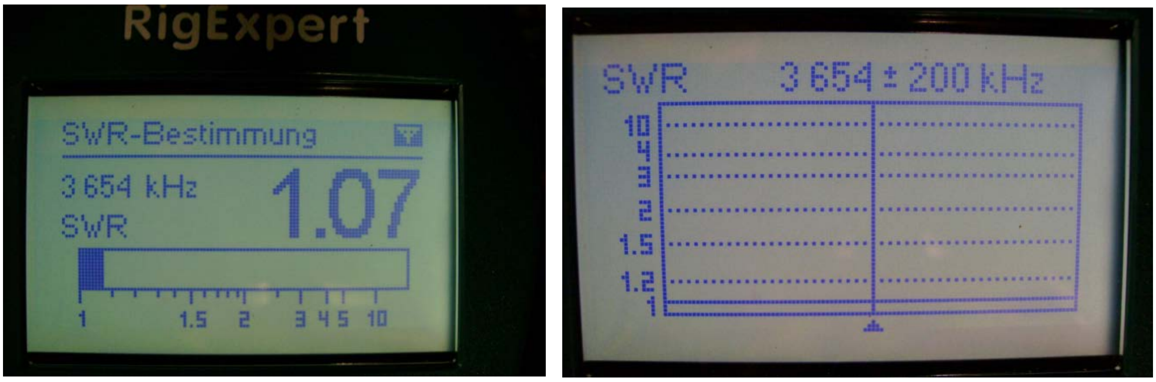


Figure 6.12 (outputs terminated with 100 ohm) SWR at 3654 +-200 kHz

6.4 50 Ohm Dummy Load

The dummy load dissipates the power that is reflected when radiators are not properly matched. The built-in dummy load can absorb short-term peaks of about 250 Watts, but constant 50-70 Watts should not be exceeded. The radiators are largely decoupled from each other by the phasing-box.

6.5 The Remote Control Box

The 12 Volt relays (16 Amps/250 Volts) in the phasing-box switch the radiators to the outputs with in the required phase for the preselected directions S, W, N and O. The built-in meter indicates the reflected power flowing into the dummy load. It is scaled in dB,
0 dB corresponds to about 100 Watts. In well-designed systems the power flowing into the dummy load should only be a few percent of the input power on the resonant frequency. A dummy load is sufficiently rated with 10%-20% of the transmitter output power.

6.6 Control Lines

An 8-conductor-cable (e.g. IT-patch-cable) transfers the following signals: reflected-power level, relay control for K1, K2 and K3, each with their own ground returns. In this configuration patch-cable can be used up to a distance of 100 meters.

6.7 Control Box Power Supply

A 12 Volt / 1Amp. power supply is required for the coupler. Switching power supplies should be checked first for causing interference in the receiver! Often the 12 Volt outlet of the transceiver can be used to power the control box.

6.8 Power Test

In Figure 6.13 the coupler is tested with 1.2 KW. Fig. 6.14 shows the black transformation lines, the yellow control lines and the dummy loads of the test setup.

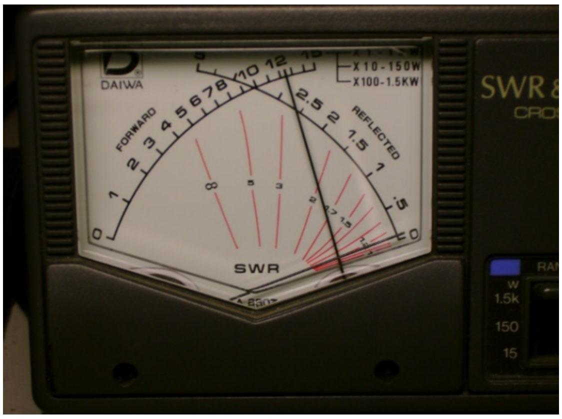
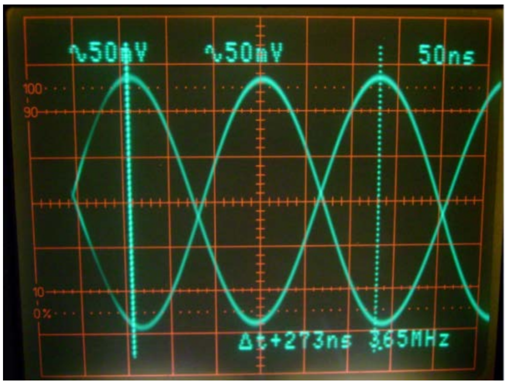


Figure 6.13

Figure 6.14

6.9 Phase Measurement of a 3.5 MHz Coupler

The Figs. 6.15 and 6.16 are screenshots of the output signals of the coupler. They prove the required conditions: the same amplitude and proper phases of the signals on the operating frequency.



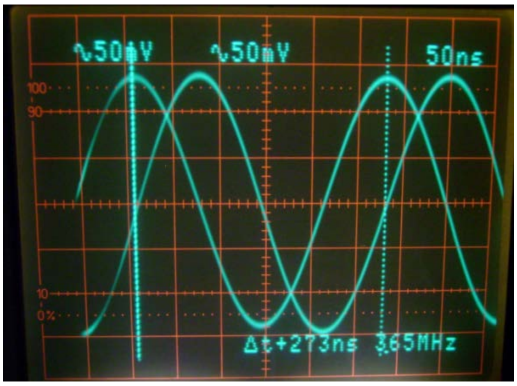


Figure 6.15 180 degree phase

Figure 6.16 90 degree phase

6.10 Complete System

Fig. 6.17 shows the complete kit consisting of the phasing box, the remote control box, the transformation lines (black), the control cable (yellow) and the power supply.

Figure 6.17

6.11 Setup Notes

Put the box in the center of the planned 4-square and lay the phase lines out in the desired directions. Keep them at an angle of 90 degrees to each other. If using PE-insulated lines the positions of the radiators are fixed by their ends, but then the spacing is not quite λ/4. With foam-insulated lines the positions of the radiators have to be chosen at a distance of 0,175 λ (0.25 λ / 1.41) from the center for a λ/4-spacing.
Before connecting the radiator to the transformation lines each radiator has to be checked for resonance on the operating frequency with a suitable analyzer.
Important: all other radiators are not connected to their feed-lines!
After connecting the feed-lines you can measure the transformed feed-point impedance of the antenna at their ends in the center of the 4-square. It should be about 100 Ohms. Values between 90 and 110 Ohms are acceptable.
To check the display on the remote control box, remove all λ/4-lines from the phasing-box; only the built-in dummy load is effective. As the output ports are open, the input power is totally reflected into the dummy load. You can check this by sending 100 watts. The instrument should then read about 0 dB.
The SWR remains low in this test!

6.12 Operation with 2 Radiators

You can also run a 2-element array with the coupler. The two radiators should be placed at a distance of approximately λ/4 from each other. When the radiators are connected to the “North” and “East” terminals, the array radiates in the direction of the element, which is connected to the port "North", if “North” is selected at the control box. When selecting the direction “East”, the array radiates in the opposite direction. The dummy load absorbs the power reflected from the radiators. Thus, the two radiators are fully decoupled. 50 ohm cable of the same length must be used as feed-lines.

6.13 4-Square Antennas for TS7N, 5A7A , VP6DX and TF7X

The 80m 4-square at TS7N consisted of four full-size aluminum verticals, each with one elevated gull-wing radial. For transportation by aircraft, a maximum length of 2 m was allowed. For this reason the home-made verticals consisted of 2 meter segments starting with 50 x 2 mm aluminum tube at the bottom ending up with 10 x 1mm.
For safety the verticals were guyed in three planes and in four directions. In Fig. 6.20 the verticals defy wind force 7-8. Fig. 6.18 shows the erecting of a prepared radiator. In Fig. 6.19 the rather fragile structure is being erected by means of an auxiliary pole. In Figure 6.21 the 80m 4-square is on the left, on the right there is a 22m vertical with a 40m wire 4-square interlaced in the guy-wires. Since wire antennas can hardly be seen on photos the sketch in Fig. 6:22 illustrates the construction principle.

Figure 6.18

Figure 6.19

Figure 6.20

Figure 6.21

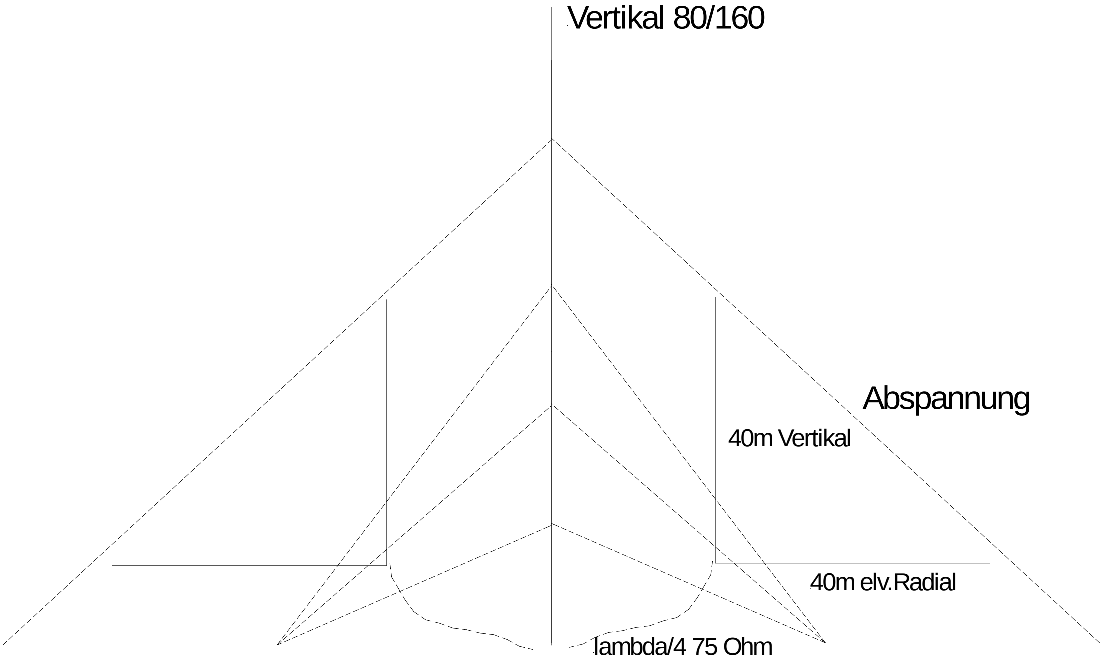


Figure 6.22

Most suitable for DX-expeditions very light 4-square antennas for 20, 30 and 40 m can be built with this construction.
In TF7X on EU71 we used such a wire-4-square for 20 and 40m.

Figure 6.23

In Fig. 6.23 the author is standing in front of the 80m 4-square antenna at 5A7A. It was his contribution to this successful DX-pedition. The array consisted of four 18m Spider-beam fiberglass poles, on which the wires for the radiators were spirally wound. The phasing-lines were made of 75 ohms PE-insulated coax with a VF of 0.66. Each vertical had four 24m ground radials soaked in salt water on the beach.

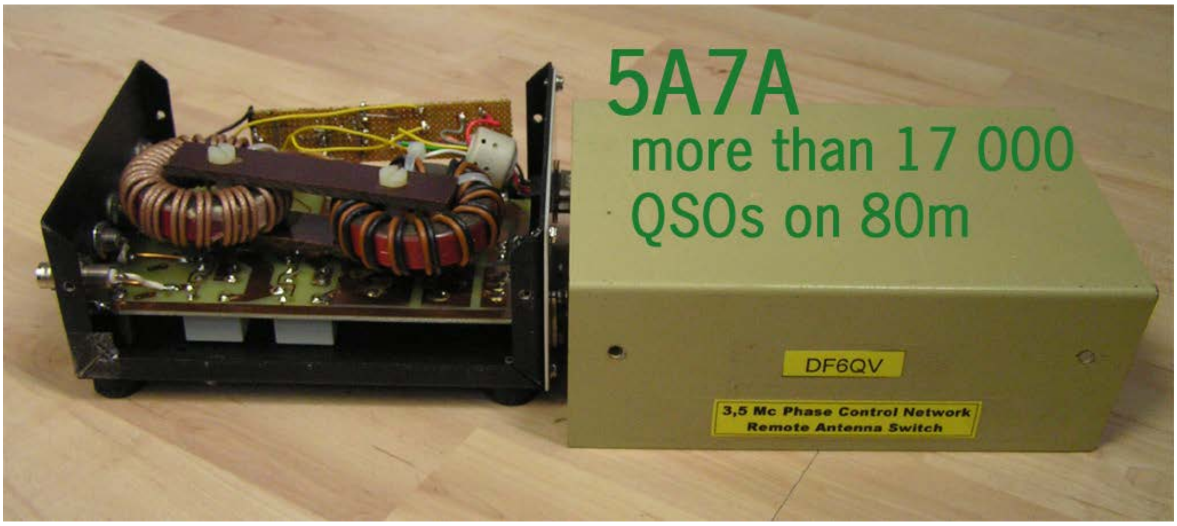


Figure 6.24

Fig. 6.24 shows the coupler for 80m that was also used at TS7N in 2003.
(cf. http://ts7n.tk/)

Figure 6.25
I had the exciting pleasure to support VP6DX with 4-square-couplers for 80m, 40m and 30m. For more details see http://ducie2008.dl1mgb.com/theteam/in dex.php

Figure 6.26

Figure 6.27

Figs.6.26 and 6.27 show a wire-4-square for 20m in Iceland on EU71 that was supported by the guy-wires of a 15m fiberglass mast as sketched in Fig.6.22. For more details see

6.14 Field Test with a 40m 4-Square

FS-measurements were carried out with a PERSEUS-RX and a CIA-HF Antenna Analyser as a remote TX.
The W-O axis showed an F/B-ratio of -16 dB and side-lobes of -15 dB and -20 dB, while rectangular to the square the rear elements showed -6 dB and -8 dB.

4-Square at DF0AT

Measurements on 19-June-2014

N -15 dB

W -16 dB **O** 0 dB ⇐ **TX**

S -20 dB

W -6 dB **N** 0 dB

⇐ **TX**

S -8 dB **O** 0 dB

W -14 dB

S -15 dB **N** 0 dB ⇐ **TX**

O -17 dB

Bild 6.28

Fig.6.29 DJ2YA (right) and DF6QV (left) checking the pattern of a 4-SQ

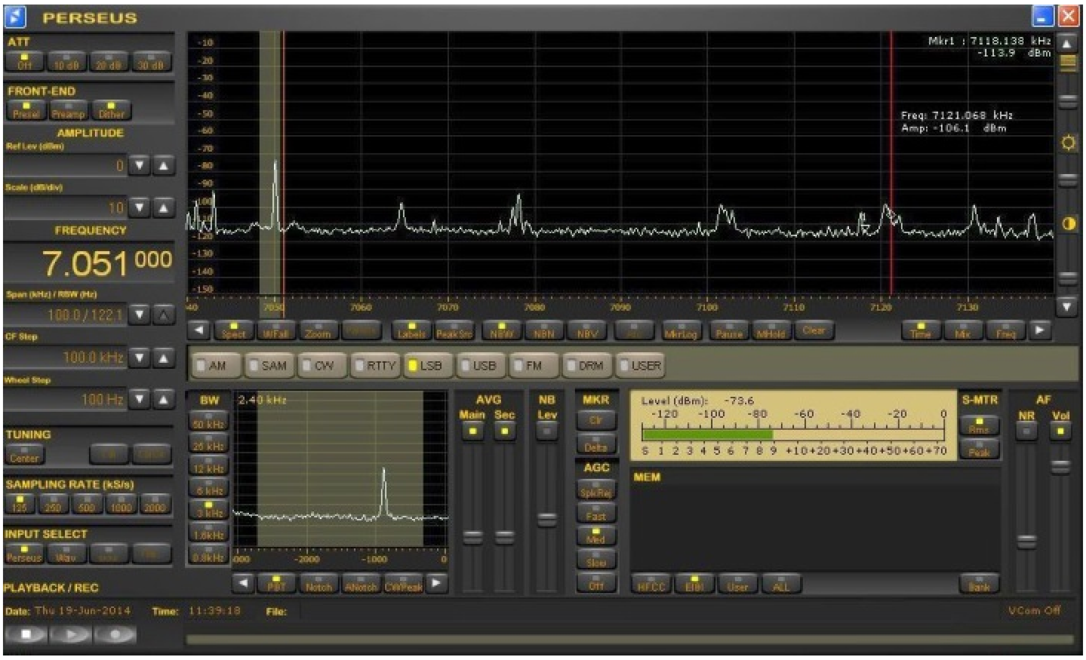


Fig. 6.30 Screenshot of the PERSEUS Test-RX

7 Literature and Web Links

QST, March 1965, “A Switchable Four-Element 80m Phased Array”

HAM RADIO, April/May 1984, “The Branchline Hybrid” Part 1+2

QST, January 1978, W2CQH, “Twisted Wire Quadrature Hybrid Directional Couplers”

HAM Radio, June 1982, W1OOP, “The Hybrid Coupler”

ON4UN, Low Band DXing, 3rd ED

MMANA MM Antenna Analyzer

EZNEC Antenna Software by W7EL
http://www.eznec.com/

Mirometals Application Notes, RF Applications by Jim Cox, 2 PDFs
http://www.micrometals.com/appnotes_index.html
http://www.micrometals.com/materials_index.html
Mini toroid calculator by DL5SWB
http://www.dl5swb.de/html/mini_ringkern-rechner.htm

OE3REB, „120 gegen 4“, pp 53 f
http://oe9.oevsv.at/export/oevsv/download/QSPArchiv/2009/QSP200910.pdf

N6LF Series of QEX articles on ground system experiments
http://www.antennasbyn6lf.com/2009/12/series-of-qex-articles-on-ground-system-experiments.html

K2AV, “Low Band Antenna Solutions”
http://www.w0uce.net/K2AVantennas.html

K3LC, “A Four-Square with Eight Directions of Fire”
http://ncjweb.com/bonus-content/k3lc4squarea.pdf

Propagation Links:
http://www.astrosurf.com/luxorion/qsl-hf-tutorial-nm7m6.htm
http://www.qsl.net/g3yrc/hf%20propagation/hf%20propagation.htm

„LTspice“ of Linear Technologies, free E-CAD Software
http://www.linear.com/designtools/software/

Fritz Dellsperger , Smith V3.10
http://www.fritz.dellsperger.net/

DF6QV, PDF “4-SQs With 90-Degree-Couplers”
http://www.rrdxa.eu/rrdxa/news-english/4-sq-antennas-by-df6qv.html

DF6QV, QRZ.com
http://qrz.com/db/DF6QV

All URLs were available on 01.05.2014 .
Credit: DF6QV unless otherwise stated