QUASIOPTICAL MULTIPLIER ARRAY WITH SEPARATE WAVEGUIDE FEED

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1 Abstract

A new type of Submillimeterwave source is presented consisting of a high power pump oscillator, a power splitting network and a varactor multiplier compound with quasioptical output. Pump power is delivered by a Russian backward wave oscillator (BWO) up to a level of 500 mW at a frequency of 145 GHz. The multiplier setup contains four strictly parallel working varactor diodes in the quadrupling mode leading to an output frequency of the SMMW–source of 580 GHz. The diode output circuit is realized quasioptically simultaneously acting as a tunable power combiner. The idler terminations are impedance tuned by a quasioptical filter network.

2 Introduction

Low conversion loss in SMMW–heterodyne receivers requires sufficient local oscillator power in order to drive hard the nonlinearity. On the one hand a combination of a semiconductor oscillator and a subsequent frequency multiplier appears to be a proper solution of a compact and flexible local oscillator source. On the other hand the multiplier output power is limited because all available diode types show saturation effects driven with some mW of input power resulting in a maximum of multiplication efficiency of SMMW–varactors in dependence of their pump power [1,2]. Besides varactor design for high cut–off frequencies on the one hand and for high power handling capability on the other hand require a tradeoff of the diode parameters [3]. Therefore a promising approach to a higher output power of compact SMMW–multiplier sources could be the distribution of input power to several parallel working multiplying elements followed by an efficient power combination of the single output signals. In our approach a multiplier containing four parallel and synchronously working varactors has been built.

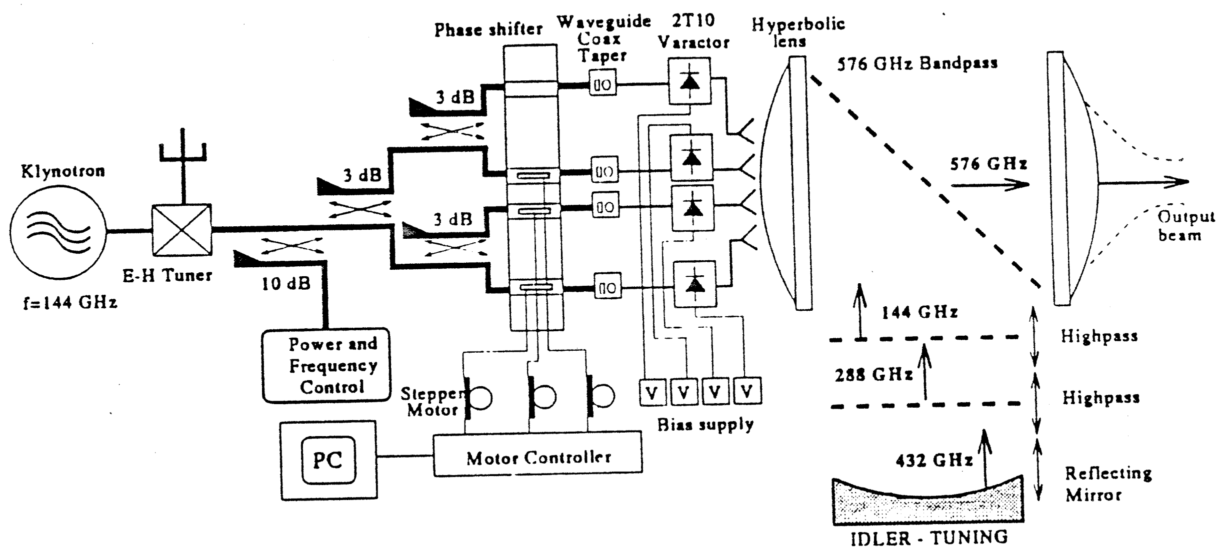

Figure 1: Principle circuit of the SMMW–source

3 General description of the source

Input pump power at the fundamental is delivered by a Russian BWO ($P_{max} = 500$ mW, 142 GHz $< f_{out} <$ 148 GHz) and divided into four phase coherent channels of equivalent power transmission by a waveguide network of three 3 dB directional couplers. This makes an independent adjustment of amplitude and phase for each fed device possible. In a computer–controlled four channel phase shifter the phase relations between all channels can be adjusted. Each diode of the multiplying array is fed separately via a semi-rigid coaxial line which operates in the fundamental mode up to a frequency of 894 GHz. Every transition from waveguide to coaxial line is tunable using a sliding contactless short [4]. The diodes (whisker contacted 2T10 UVA [1] varactors with honeycomb structure) are placed at the end of the inner coaxial conductor. The multiplying efficiency of each diode is controlled by a separate bias feed. The whiskers act as a quasioptical long–wire antenna array in the output circuit. The antennas are lined up in form of a broadside array with the main lobe perpendicular to the linear array geometry to avoid shielding effects. They are placed in front of a movable plane mirror to improve the directivity of radiation. Optimum power combining is achieved for a distance of 0.2 mm between the antennas and 0.3 mm spacing between array and reflector. Power combination can also be controlled by quasioptical interference effects which are controlled by the relative phase shift of the pump signals. The output beam is collimated using a hyperbolic lens; the fourth harmonic at 576 GHz is filtered by a bandpass (i.e. a resonant metallic mesh) and couples through a second lens into a quasioptical beam waveguide which finally leads to a cornercube mixer. The idler circuits at the second and third harmonic and the fundamental circuit are tuned by a quasioptical network consisting of two movable highpass filters and a movable mirror. This offers the possibility of an independent impedance termination adjustment for each

[1] UVA : University of Virginia

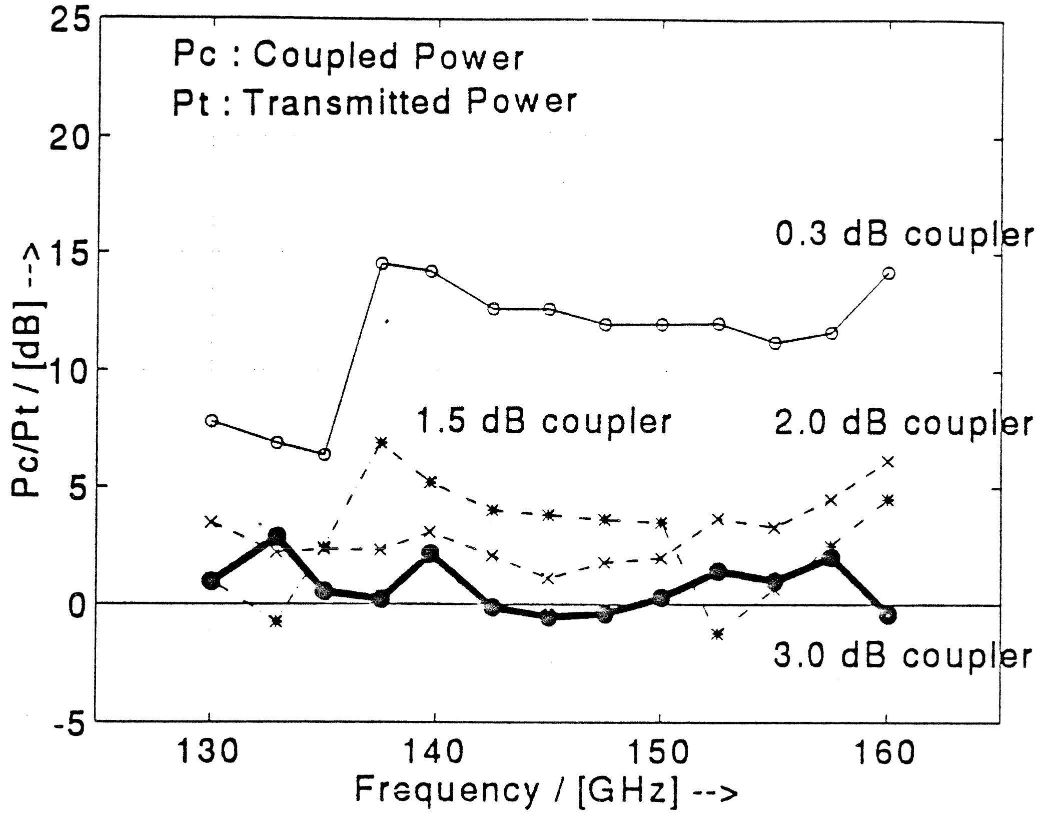

Figure 2: Measured power splitting of different directional couplers

frequency circuit.

4 Power–splitting network

Parallel and optimum operation of each diode requires separate pump signal feed. This is realized by a power-splitting WR-6 hollow waveguide network (1.651 mm * 0.8255 mm). The source is matched to the waveguide network by an E–H tuner. Feeding with a single high power source secures phase-coherent input signals for all active diode elements. Phase shift between the channels is controlled by a four channel phase shifter. Power is splitted into four channels with equal amplitude by 3 dB directional couplers. Phase shifter, E–H tuner and 3 dB directional couplers were developed and machined in our laboratories and are discussed in detail subsequently.

4.1 3 dB directional coupler

The directional couplers are realized as top wall couplers with circular holes [9] etched in a 12 μm thick, gold plated brass foil. For the 3 dB types we chose two perforation lines having Tschebyscheff design with hole diameters between 250 μm and 448 μm.

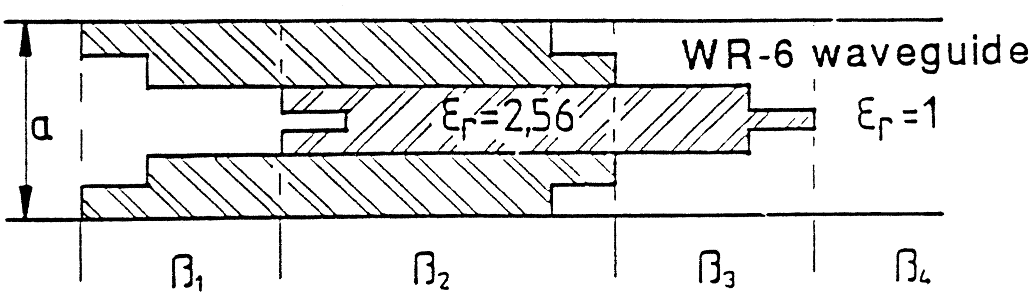

Figure 3: Schematic of four-channel phase shifter

Measurements of different couplers are shown in Fig.2. Here the relation between the power amounts at the two different output ports is displayed as a function of the input frequency. A coupling coefficient variation lower than 1 dB was achieved within a bandwidth of 15 GHz around the center frequency. The couplers were designed for an operating frequency of 145 GHz transmission loss being below 2 dB.

4.2 Four channel phase shifter

In principle the four feeding channels are designed to be of equal length. Manufacturing may cause deviations from the standard electrical length of the pump signal channels. For compensation of these deviations we constructed a four channel phase shifter making a relative phase shift between the different signal paths possible. Consequently, every defined phase relation between the single pump signals of the diodes can be adjusted according to an optimum power combining between the different antenna signals in the output circuit. Another feature is the possibility of an output beam steering (*Phased array*).

Phase shifting is achieved by moving dielectric stripes made of polystyrene ($\epsilon_r = 2.56$) inside the WR-6 waveguide. Every waveguide contains three dielectric stripes. The outer ones are fixed and a central stripe can be reproducibly moved relative to them by a computer controlled stepper motor. For matching dielectric filled waveguide and hollow waveguide, the stripes are equipped at their ends with $\frac{\lambda}{4}$ - transformers.

Generally the phase shifting structure consists of four waveguide sections with different phase constants β_i. Moving of the central stripe corresponds to a variation of the single line lengths or of the resulting phase constant of all sections respectively. A high similarity between the transmission characteristic of the different channels was realized. A highly linear phase shift was measured from 0 to 200 degrees in dependence of the stripe displacement (Fig.4). Transmission loss of each channel sums up to 3.5 dB.

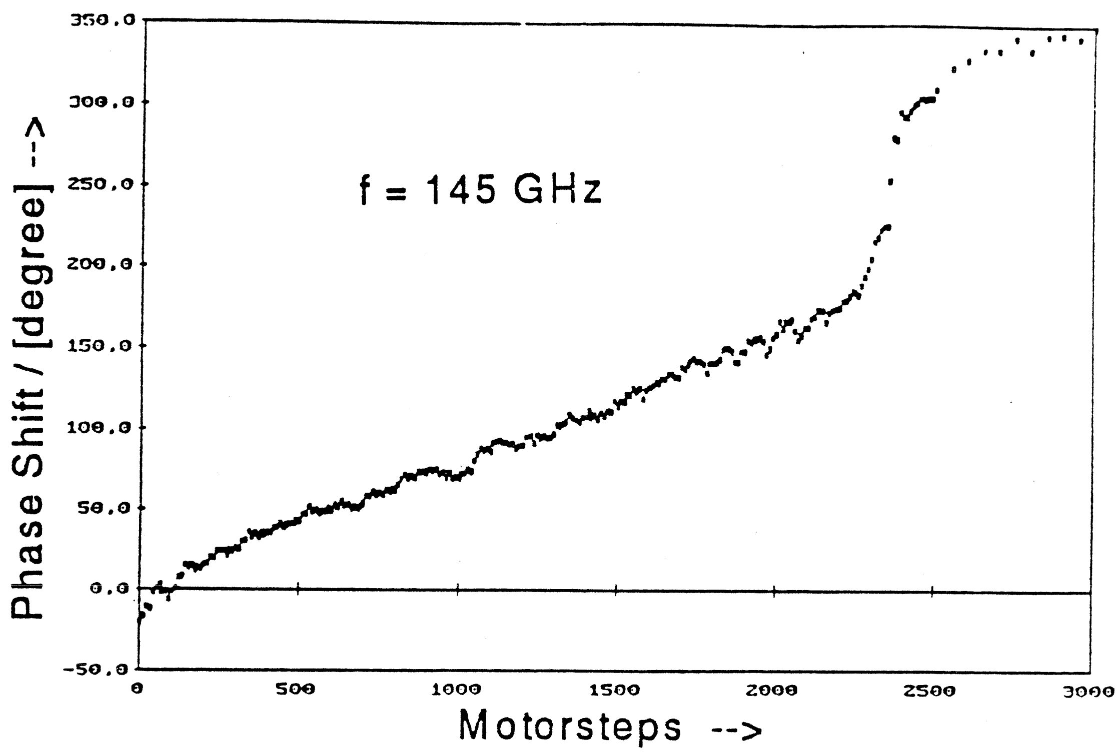

Figure 4: Measured phase shift

4.3 E–H tuner

The E–H tuner was also designed in our institute. For matching operation we use non-contacting shorts with cylindrical cross-section. The micrometer driven plunger contains seven sections with alternating characteristic impedance Z_l. We realized plunger diameters of 0.8 mm and 0.25 mm what corresponds to a Z_l of approximately 1.88 Ω and 71.7 Ω.

This leads to an impedance drop by a factor of $7.6 * 10^{-9}$ for a given operating frequency of 145 GHz [4]. We measured a tuning-dynamic of more than 24 dB over a bandwidth of 10 GHz around a center frequency of 145 GHz as illustrated in Fig.5. The transmission loss of the device is lower than 1dB.

5 Coaxial–waveguide transition

Nearly all MMW- and SMMW multiplier setups contain a coaxial waveguide transition because the diode is usually mounted on a coaxial choke structure. The other end of the inner conductor is usually soldered with the waveguide bottom in order to close the bias circuit of the diode. Similarly the electric field of the H_{10}-mode in the rectangular waveguide is shortened at its maximum field strength what negatively affects the coupling efficiency from H_{10} - waveguide to TEM - coaxial mode. Therefore, we developed a new kind of transition which simultaneously combines a closed bias circuit and a more efficient mode conversion. In our model the coaxial line is elongated through the lower waveguide wall and is shortened at a distance of a quarter-wavelength away from the waveguide

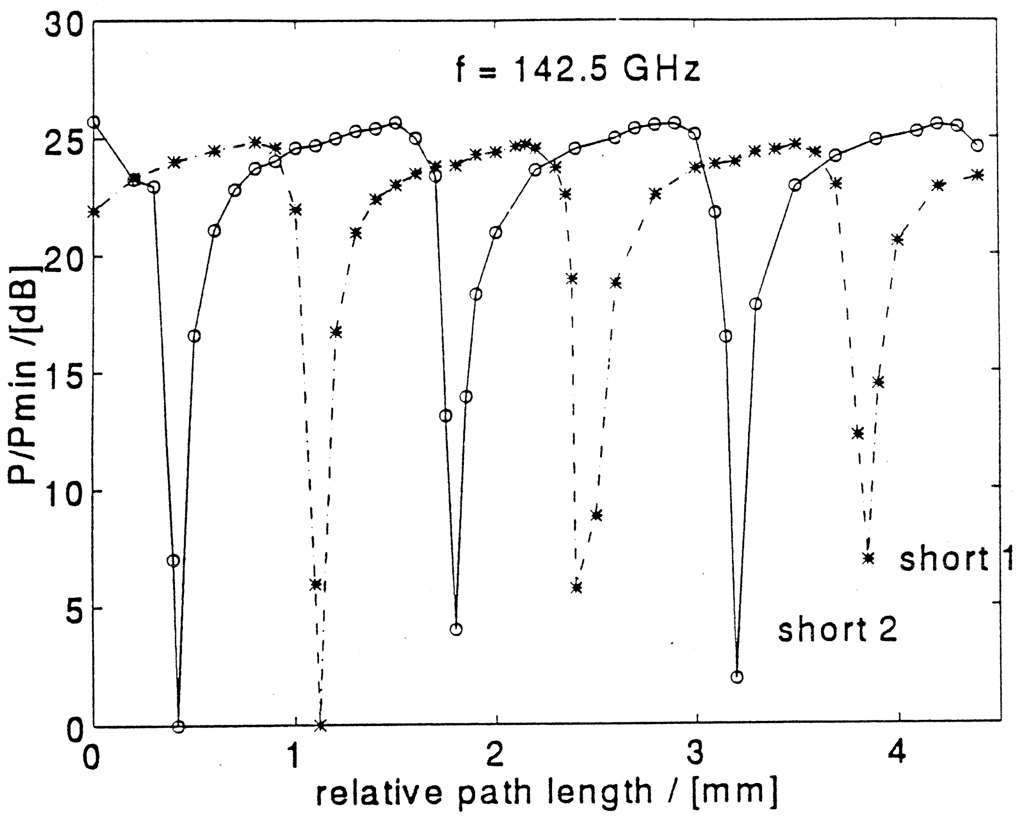

Figure 5: Reflection measurement of the E–H tuner

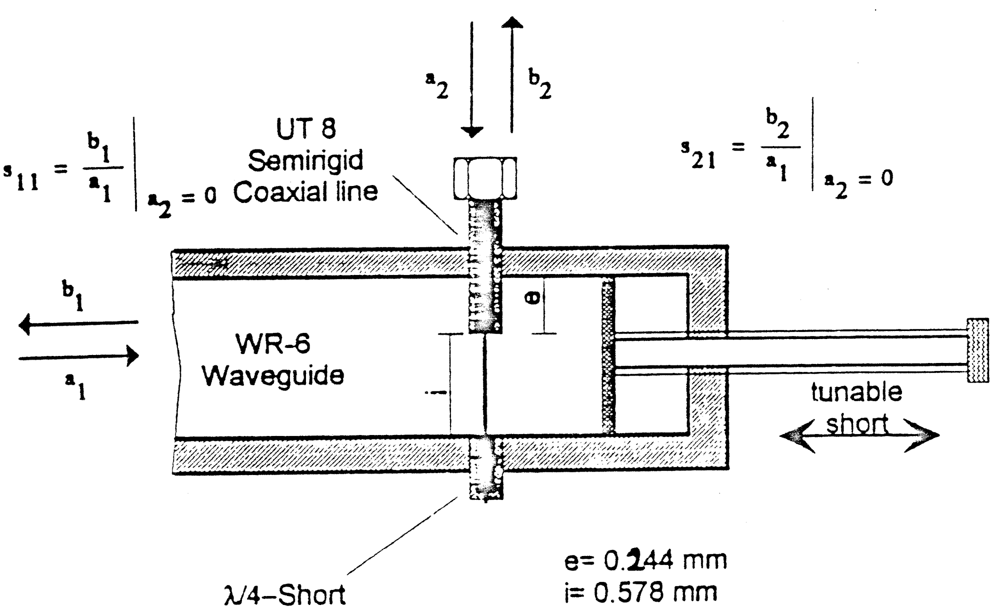

Figure 6: Schematic of waveguide–coaxial transition

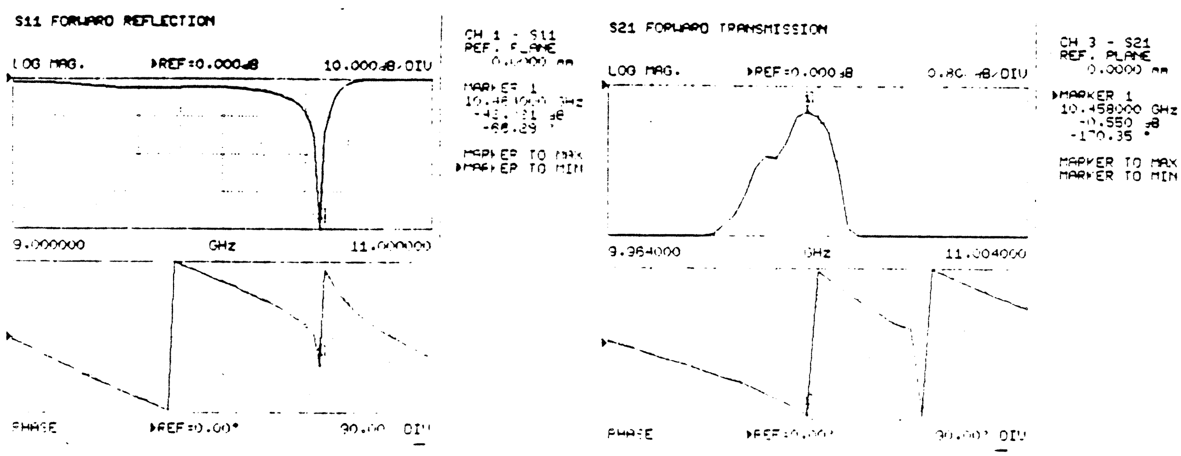

Figure 7: S–parameter measurement of waveguide–coaxial transition

bottom. So, the short is transformed approximately in an open circuit at the waveguide bottom and simultaneously the bias circuit is closed. Similarly the outer conductor of the coaxial line dips into the waveguide at a length of 0.244 mm . The transition from WR-6 waveguide (1.651 mm * 0.8255 mm) to the semi–rigid coaxial line (d_i=0.05 mm, d_a=0.1 mm) is shown in fig.6. The axial line profile in the transition region was realized using a photolitographic etching process. The dielectric is left standing in the area of the open laid inner conductor so as to achieve a higher coupling–bandwidth of the transition. The transition can be adjusted with a non contacting movable short. It was tested by measurements at X–band in a frequency range of 9 to 11 GHz as illustrated in fig.7. Here a transmission loss improvement of about 4 dB compared with the standard design was achieved (compare Table 1).

	Bandwidth	Power coupling
Standard	4.82 GHz	36.2 %
New transition	5.75 GHz	88.1 %

6 Quasioptical multiplier array

The varactor array is fed by four separate semi–rigid coaxial lines of about 1 cm length. They are monomode up to 894 GHz and have a transmission loss of about 1 dB/cm. The diodes (chipsize 100 μm * 100 μm * 100 μm) are soldered onto the end of the inner coaxial conductor. For mechanical stability the inner conductor is enlarged to a diameter of 150 μm by soldering a 12 μm thin brass bush onto its end. The semi rigid lines are bent as illustrated in fig.8 to realize a distance of 0.2 mm between the active elements. The lines are glued with epoxide resin inside the waveguide block to achieve high mechanical stability. As active elements we use 2T10 UVA honeycomb varactors (C_{jo} = 5.5 fF, R_s = 9 Ω) which are contacted with a 4λ whisker at 576 GHz. The coaxial lines were passed through an oblong hole drilled in a gold plated and polished brass plate. Consequently, the whiskers correspond to a long–wire antenna array on a reflecting ground plane. Every whisker is contacted and manipulated by its own adjuster which is simultaneously used to provide

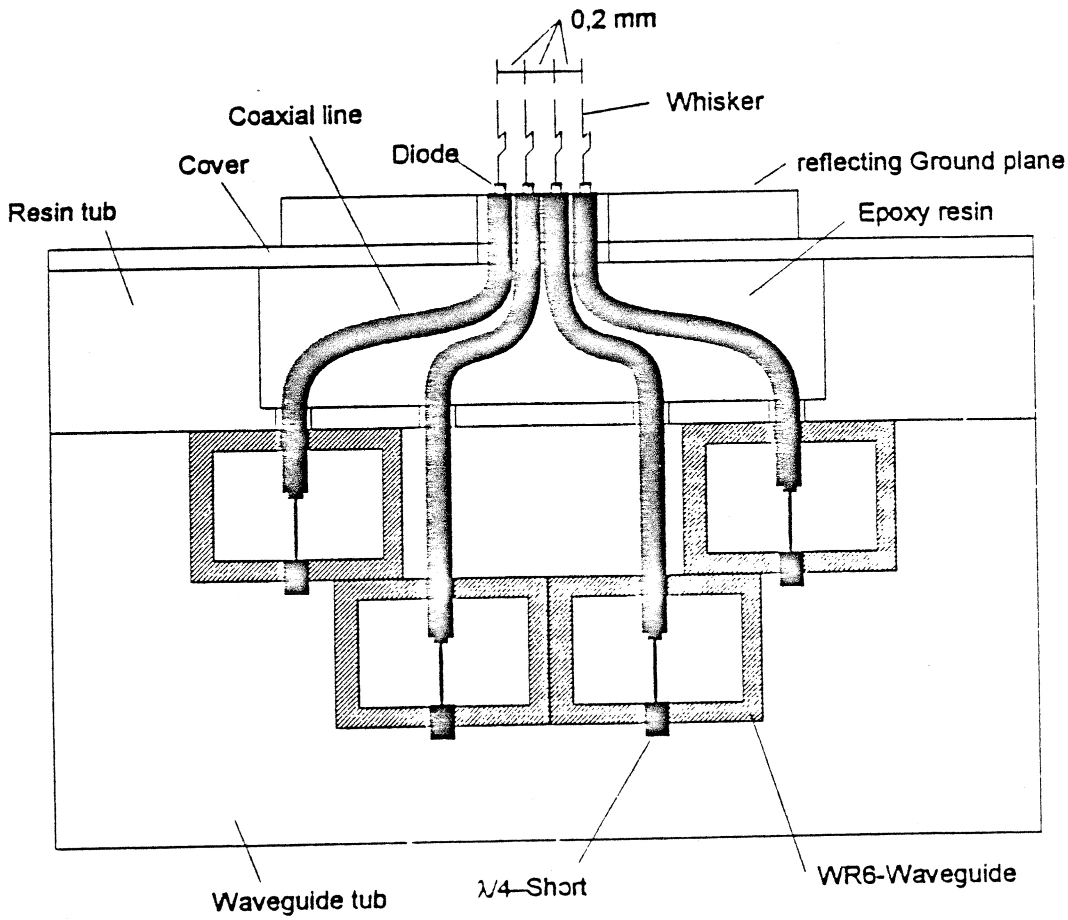

Figure 8: Schematic of coaxial feeding network

each diode with a seperate bias. So, a suitable impedance level for each diode performing optimum multiplying efficiency can be selected. For a higher directivity of radiation the array is positioned in front of a plane reflector. This reflector is micrometer-driven to adjust an optimum power combining in the output circuit. Choosing this configuration the array results in four active and twelve virtual radiation elements. With the phase shifter integrated in the power splitting network we can enforce an in-phase steering of the single radiating elements leading to a 'broadside array' configuration. Consequently at the output all active elements are working strictly parallel because their main lobe is perpendicular to the linear array geometry avoiding any shielding effects between the active devices.

The array geometry was optimized according to power combining in the far field. The array pattern as illustrated in fig.11 was calculated from the electromagnetic vector potential and geometrical parameters as the distance between array and reflector as well as the distance between the single antennas were varied so as to get a high beam efficiency and a high sidelobe attenuation. The beam efficiency is hereby defined as the amount of power collimated in the main lobe related to the whole output power of the array. Optimum

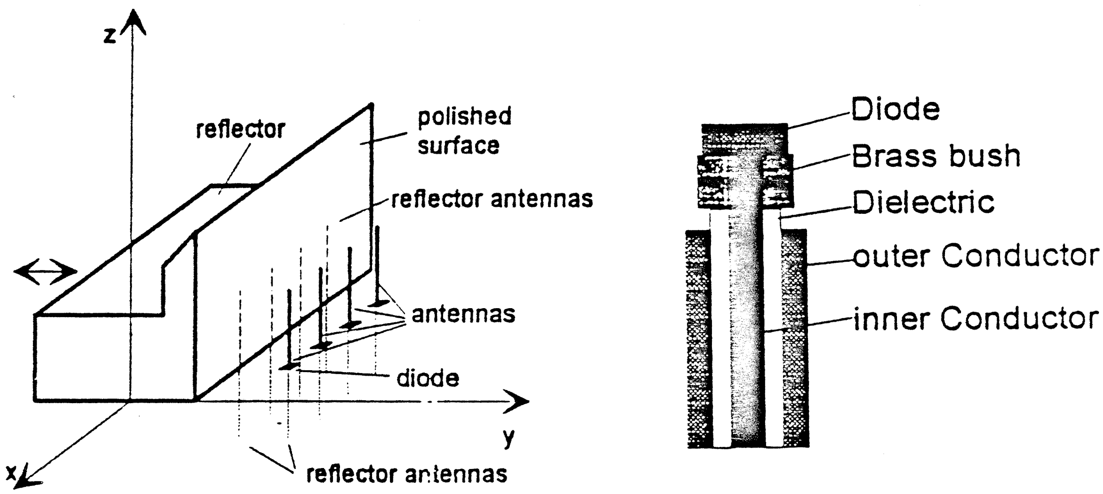

Figure 9: Schematic of diode array and diode mount

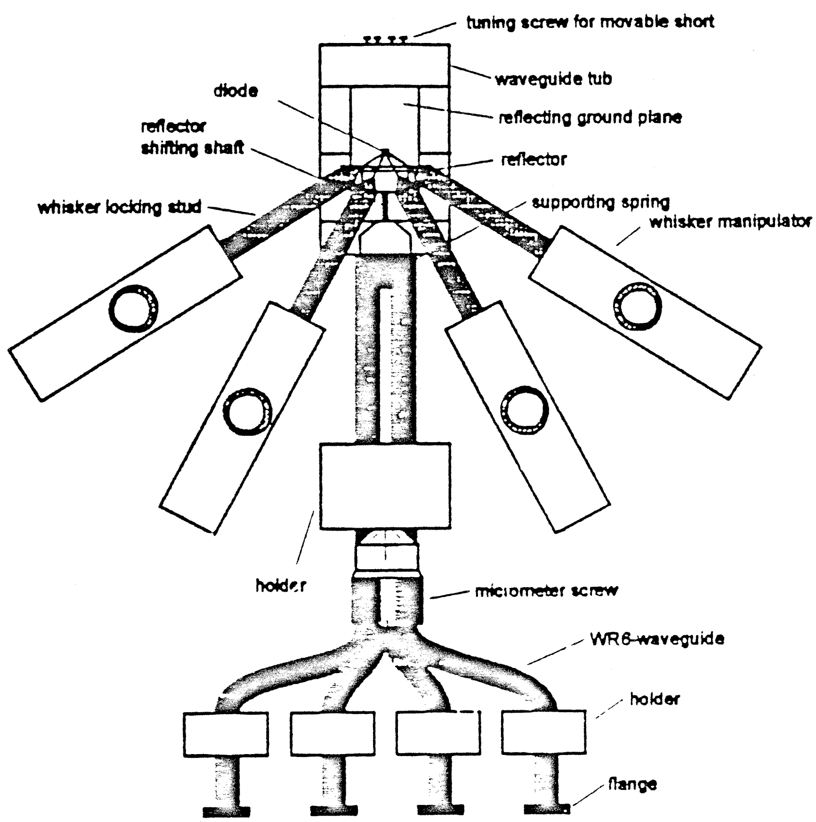

Figure 10: Setup of reflector and whisker manipulation

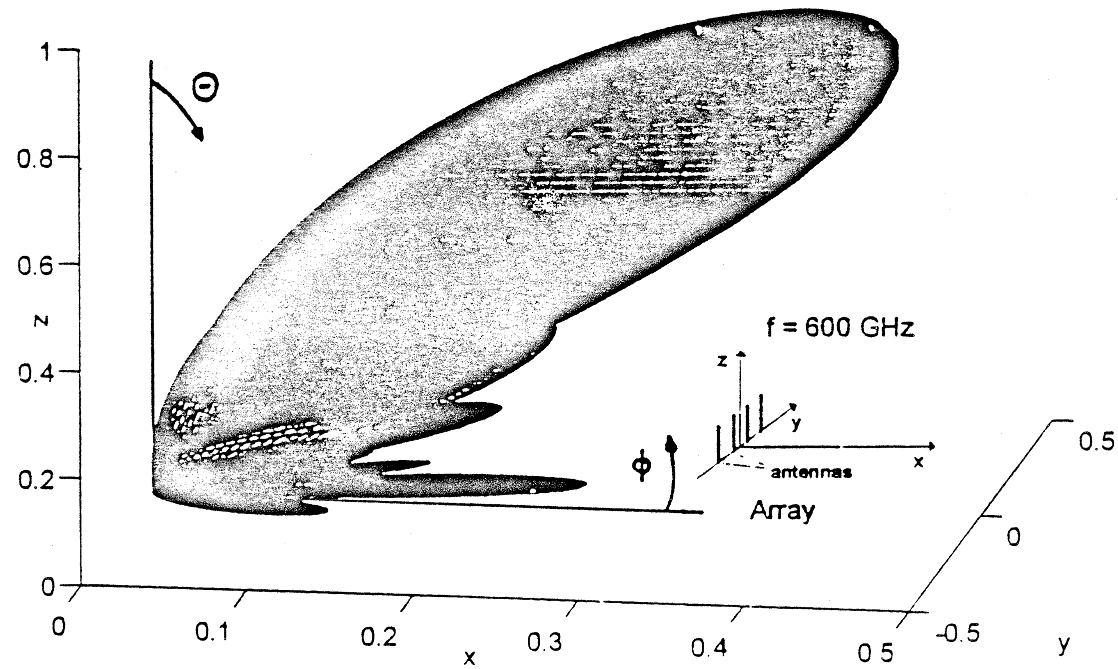

Figure 11: Beam pattern of the long-wire antenna array

values for our configuration are a beam efficiency of 89.8% and a side lobe attenuation of 9.5 dB. These optimums are coupled with a reflector distance from the array of 0.3 mm and an antenna distance of 0.2 mm.

Optimum beam coupling of the output pattern at 576 GHz to a fundamental Gaussian mode was depicted in Fig.12 in dependence of the distance between array center and coupling position as well as of the far field divergence angle Θ of the Gaussian beam. In the far field the coupling is nearly independent from position as expected. The output pattern at 576 GHz couples best to a Gaussian beam with a far field angle of 13 degrees what corresponds to a beam waist of 1.4 λ. According to these beam parameters the lenses of the quasioptical output network are designed.

7 Quasioptical filter network

This quasioptical network has to fulfil two different aspects mainly: At first the output power at 576 GHz has to be coupled effectively to the load, in our case a corner cube mixer; secondly the input circuit and the idler terminations have to be adjusted to a certain impedance level in order to realize a low conversion loss of the quadrupling process. This is done by quasioptical bandpass filters, highpass filters and mirrors as shown in fig.13 . Several realizations of a quasioptical filter network are possible. The main advantages of the presented setup are the negligible attenuation of output power by idler filters and the possibility of separate impedance tuning for each frequency circuit.

The principle of the output filter network works as follows: The first hyperbolic lens

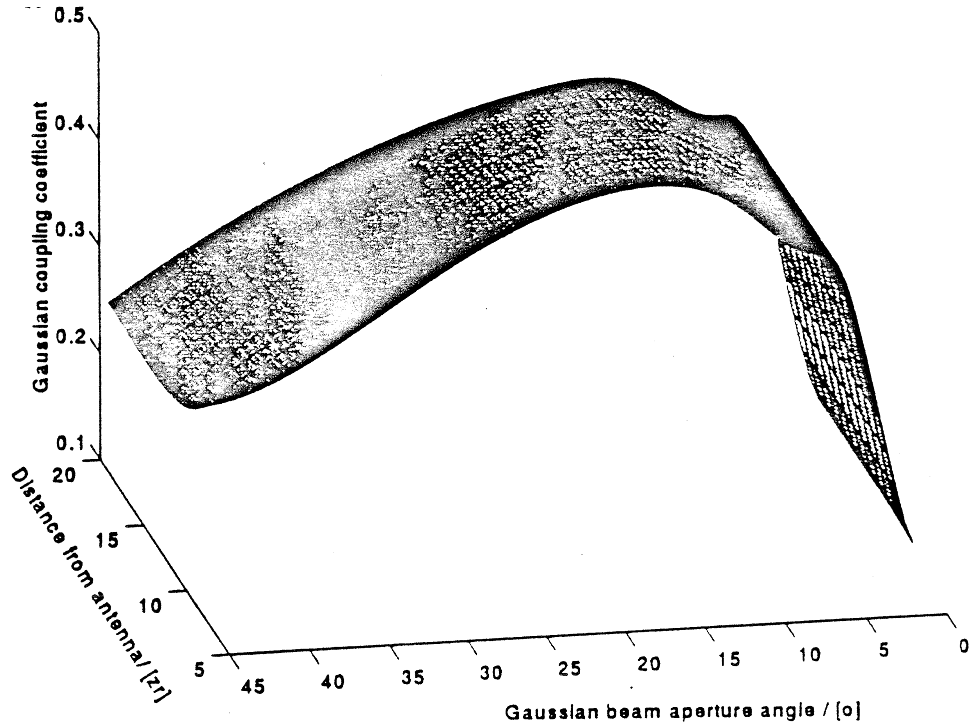

Figure 12: Gaussian beam coupling of the array

makes a transformation to a Gaussian beam of nearly plane phase surfaces. The bandpass filter is placed under an angle of 45° to the output beam. Consequently the power amount at 576 GHz is separated from the electromagnetic power of the other frequency circuits at 144, 288 (2 x 144), 432 (3 x 144), 720 GHz (5 x 144) and so on. The desired signal is matched by a second hyperbolic lens to the pattern of our corner cube mixer; the input and idler frequencies are reflected onto a cascade consisting of quasioptical highpass filters and a final mirror. The first filter is transparent for frequencies above 170 GHz, the second above 320 GHz (comp. Fig. 16b). Hereby a theoretical transmission of 50 % is assumed as frontier between reflection and transmission band. Consequently the impedance termination of the input frequency circuit can be adjusted by axial movement of the first highpass filter corresponding to a variation of the reflection resonator realized by the shorted quasioptical line. Considering the antenna as coupling link between diode and resonator (Fig. 14) the reflection coefficient at the diode output behaves like

$$r = \frac{s_{11} - det\ S\ r_a\ e^{-2\gamma l}}{1 - s_{22}\ r_a\ e^{-2\gamma l}} \tag{1}$$

Hereby **S** describes the scattering matrix of the antenna and r_a the reflection coefficient of the highpass filter.

The second harmonic at 288 GHz is seperately adjustable with the second highpass. This principle of the highpass filter cascade is extendable to all other higher idler circuits but because of their with higher order strongly decreasing power content and lower significance for the quadrupling process we only provided one variable idler termination in form of a mirror for all harmonics above 432 GHz.

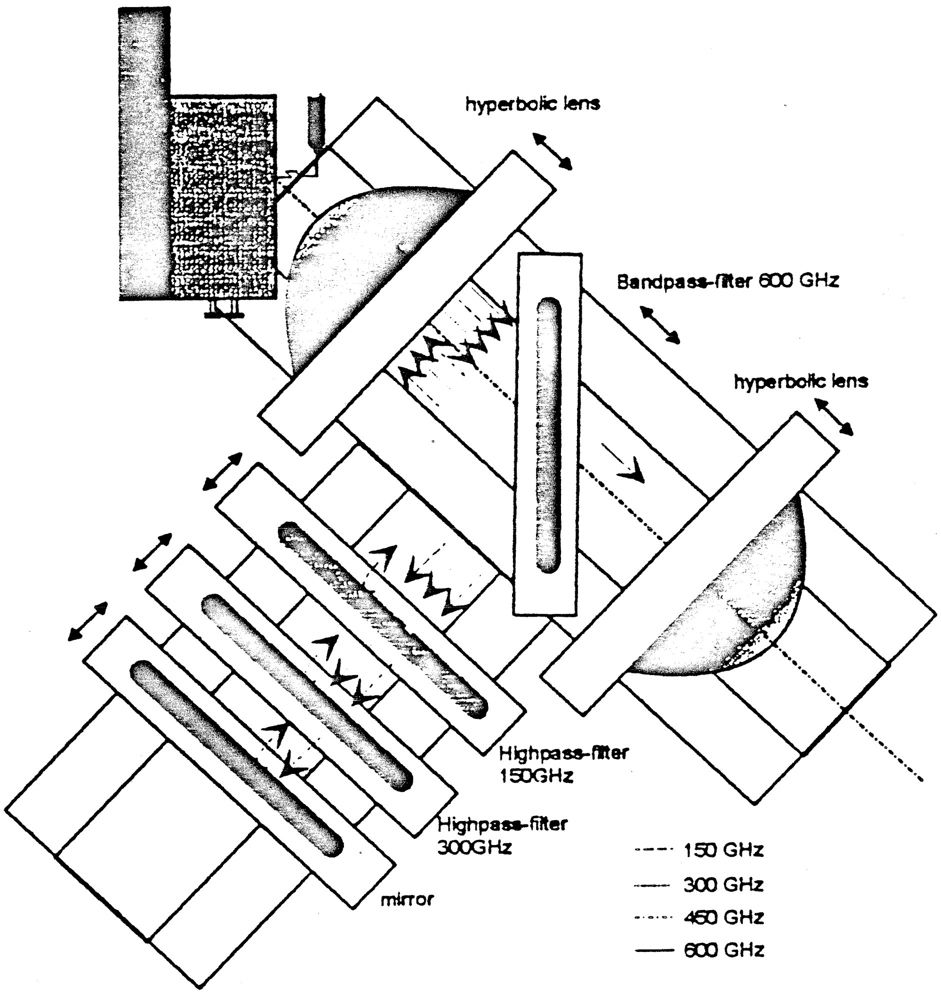

Figure 13: Schematic of the quasioptical filter network. The operating frequencies of 144, 288, 432 and 576 GHz are rounded off to values of 150, 300, 450 and 600 GHz

All components of the filter network can be swivelled independently in two orthogonal directions to make an optimum adjustment to the Gaussian beam profile possible. The edge taper of all components is greater than 35 dB to avoid diffraction loss in the beam waveguide.
For our lenses we measured a transmission loss of 0.9 dB slightly exeeding our calculated value of 0.3 dB.

The functionality of our quasioptical filter network was measured at frequencies of 144 and 288 GHz where signal generators were available. Instead of the multifrequency long-wire antenna array the quasioptical circuit was fed by test oscillator signals at discrete frequencies and the impendance tuning of the single filters was tested by measuring the power reflection coefficient at the input port of the quasioptical network. The measurements are

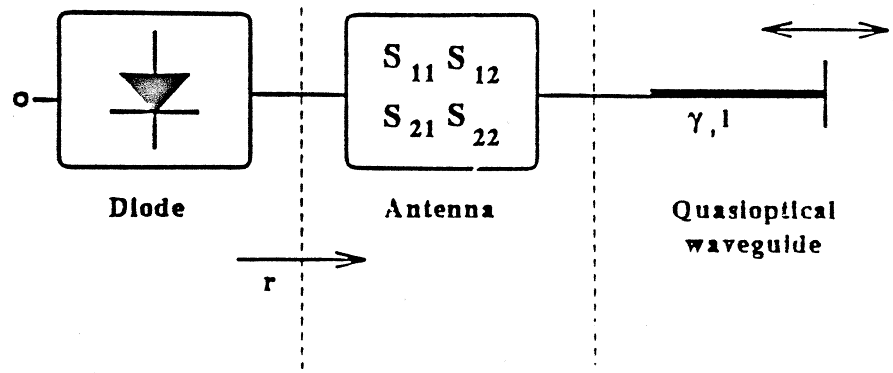

Figure 14: Equivalent network for a single idler circuit

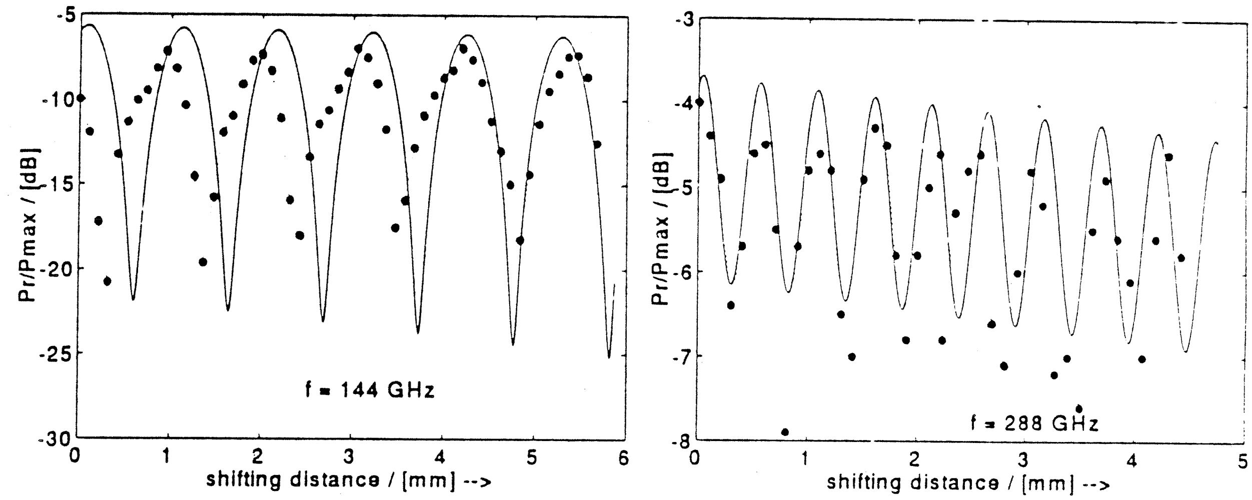

Figure 15: Measurement of quasioptical impedance tuning, Pr : Reflected power

compared with a calculation according to Eq. 1. Hereby the unknown S–parameters of the antennas are fitted to a good agreement with our check points.

The results are presented in Fig.15. The power reflection coefficient is a periodic function of the axial filter position. Minima and maxima repeat at a displacement range of half a wavelength at the test frequency. A dynamic of the power reflection coefficient of 16 dB at 144 GHz and 5 dB at 288 GHz was realized with the corresponding highpass filters. The resistive losses in our idler circuits sum up to 5.5 dB at 144 GHz and 4 dB at 288 GHz respectively. These are due to not exactly plane phase surface of the electromagnetic wave resulting in geometrical walk–off of the reflected wave, to losses of lenses and meshes and to a imperfect reflection characteristic of the high pass filters. The interference between the idler circuits is lower than 0.5 dB meaning that the tuning filter for the first harmonic has a negligible influence at 288 and 432 GHz and the other way round.

8 Quasioptical filters

As frequency selective devices different types of metallic meshes are used. They differ in their apertures according to their filter characteristic. The high–passes consist of quadratic holes in a self–supporting brass structure.
The transmission characteristic of these inductive meshes depends on the lattice constant, the stripwidth, and the thickness of the brass foil. After a photolithographic process the holes are etched into a 70 μm thick copper foil being gold–plated for a reduction of ohmic losses. The transmission characteristic of the meshes was calculated according to [6]. For the first mesh (g = 710 μm, $\frac{a}{g} = 0.79$) we realized a reflection of 81 % at 144 GHz and a transmission of 91 % at 288 GHz. For the second filter (g= 410 μ m, $\frac{a}{g} = 0.79$) a reflection of 75% at 288 GHz was reached.

The bandpass design shows cross–shaped apertures etched in brass foil of 12 μm thickness. These are equivalent to a cascade of an inductive mesh as mentioned above and a lowpass consisting of metallized patches on a dielectric substrate. The shape of the crosses is not symmetrical because it has to be placed under an angle of 45° into the beam guide. The theoretical transmission characteristic was calculated according to [7,8]. A high agreement between theory and measurement was obtained and the transmission loss for the desired signal is below 1 dB. Stop band attenuation for all powerful idler frequencies is higher than 20 dB what guarantees an efficient separation from the output mixer path.

9 Measurements

First measurements on the multiplier structure were carried out in order to test the operation of the device. Output power at 579 GHz was detected using a corner cube mixer and was displayed by a spectrum analyzer.

At first the symmetry of the power splitting network was tested. The deviation of an equal power distribution to the four channels was below 1 dB so that all diodes were assumed to be fed with nearly the same amount of input power.
In the beginning all channels of the source were tested separately. For each single channel a set of optimum parameters - plunger position and bias - was adjusted according to a maximum of the output signal. Subsequently the power splitting network was connected to the multiplier and the source was aligned to a maximum of output power by shifting the relative phase between the channels and slight corrections of the bias supply. Hereby bias voltages for the inner diodes of about -0.9 V and for the outer ones of 0.2 V guarantee an optimum quadrupling performance.
In the following the quasioptical output circuit was tuned due to a further maximization of the output. Measurement of output power as a function of the distance between the movable reflector and the antenna array shows a global maximum at a distance of 0.3 mm (Fig. 16a) which is consistent with our calculations predicting an optimum quasioptical power combining at this geometry [5].

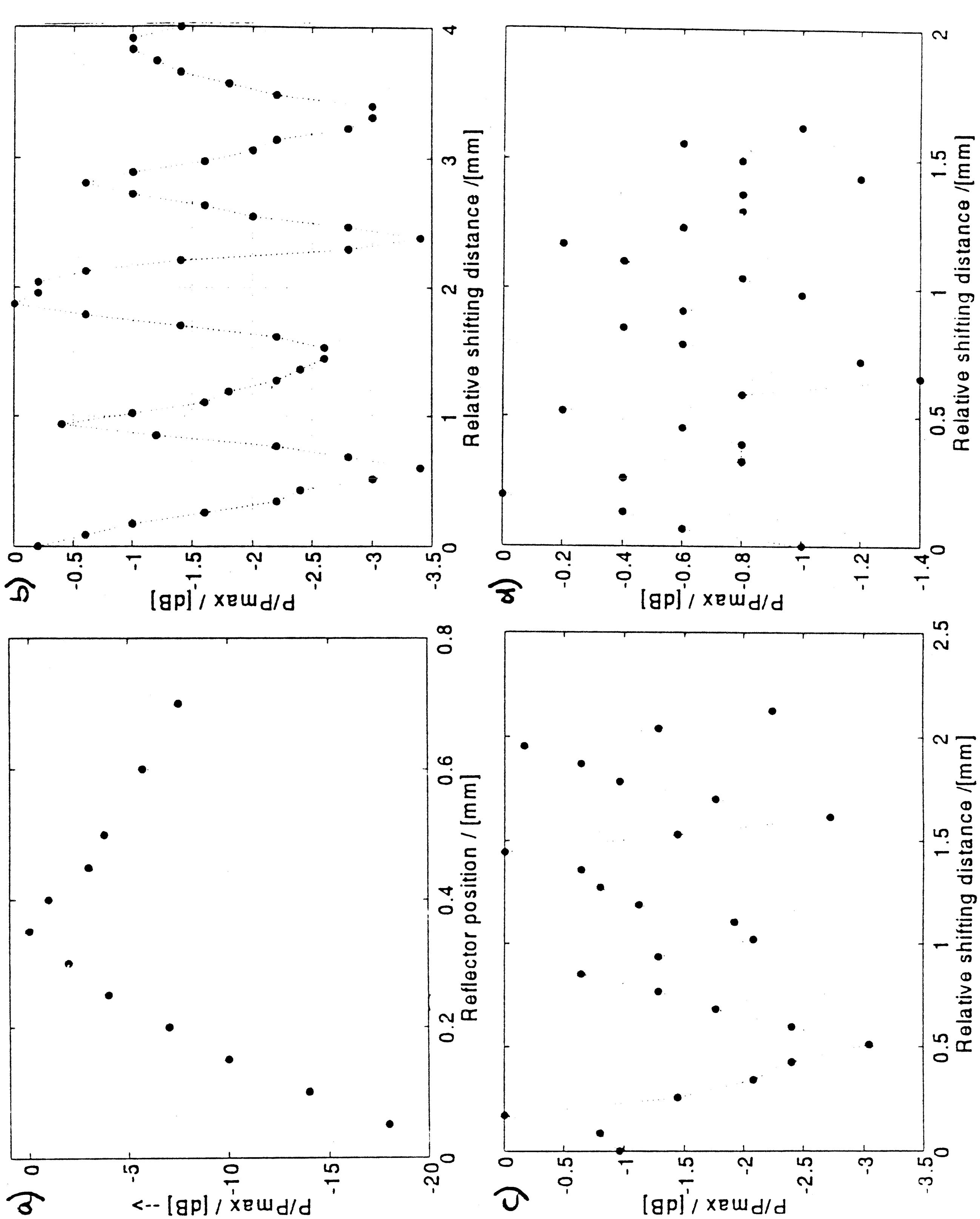

Figure 16: Measurements of output power dynamic as function of a) distance between array and plane reflector, b) Shifting distance of the first inductive mesh, c)Shifting distance of the second inductive mesh, d)Shifting distance of the mirror

An adjustment of the quasioptical filter network contains an output power dynamic of 3.5 dB in dependence of the impedance termination at the first harmonic (144.7 GHz). It is illustrated as function of the filter position in Fig. 16b. The periodicy of the output power characteristic is about 1mm what corresponds to half a wavelength of the pump frequency.
Tuning of the second harmonic with the other highpass filter shows an output power dynamic of 3.1 dB (Fig. 16c). The influence of the mirror reflecting the power amounts of all other higher harmonics is small. Here we measured a dynamic of about 1.3 dB and a periodicity at a shifting distance of 350 μm due to the third harmonic at 435 GHz. This periodicity is slightly disturbed pointing out fractions of power at other harmonics also being reflected (Fig. 16d).

Finally a maximum output power of 60 μW at quadrupling to 579 GHz was achieved due to a pump power of 280 mW. Compared with a single channel operation of the source where the whole input power was fed to one single channel and the other channels were terminated without reflection we stated an increase in output power of 4.5 dB. In this first laboratory setup we have to manage with one high power oscillator and a lossy power splitting network. Therefore it is probable that the input energy which effectively couples into each diode is to low for an optimum quadrupling performance. A multiplier setup with an integrated phase shifter in each channel driven by his own Gunn–oscillator and phase locked by a central Gunn–oscillator may guarantee a more effective diode fed and promises a further increase in output power compared with single diode operation.

10 References

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