

Wideband Fixed-Tuned Millimeter and Submillimeter-Wave Frequency Multipliers

N. R. Erickson
Millitech Corp.
PO Box 109
S. Deerfield, MA 01373

Abstract

Varactor frequency multipliers have been built with fixed mechanical tuning, which cover a fractional bandwidth >25%. These multipliers include balanced doublers and single diode triplers, with the tripler output frequency tuning up to 730 GHz. The efficiency of such a fixed tuned tripler in the 250-320 GHz band, using new design and fabrication techniques, is comparable to that of previous mechanically tuned devices. A doubler design for 200-240 GHz is not so well characterized but also appears comparable to previous tunable models.

Introduction

Millimeter and submillimeter wave frequency multipliers are required for a wide variety of applications ranging from laboratory spectroscopy to remotely operated receiver LO's on radio telescopes. While the ability of the operator to interact with the multiplier varies with the application, in every case the use of the multiplier would be simpler if there were no need to mechanically tune the device. In addition, for applications requiring considerable tuning on a frequent basis the backshorts within the multipliers can wear out relatively quickly, diminishing the power or even shorting out the waveguide with metal chips. For all of these reasons, it is highly desirable to completely eliminate mechanical tuners.

From a historical perspective, mechanical tuners are a sign of immaturity of a technology. In many microwave devices (such as waveguide mixers), tuners were initially required and have been eliminated over time. In other cases (such as mm amplifiers) tuners were never even considered as an option, and the devices have always been required to operate fixed tuned, regardless of the limitations this imposed. In the case of multipliers, tuners have many times been required to meet user expectations of power and bandwidth, but recent advances in circuit design are rapidly eliminating this need.

Bandwidth limitations of varactor multipliers

Varactor diodes are simply nonlinear capacitors having no inherent resistance except for an entirely undesirable series resistance. They develop their useful resistive component through a coupling of the input circuit to the output via the frequency conversion process. An ideal varactor (with $R_s = 0$), measured with a small test signal, thus will reflect all of its input power. Even under the case of a large input signal, varactors remain highly reactive loads, with a typical input Q around 5. This reactance produces inherent limitations to the bandwidth possible without significant mismatch.

Practical considerations to the realization of circuits also play an important role in the achievable bandwidth. Most circuits are built using coaxial or stripline circuits and in both cases, it is difficult to fabricate matching elements with very high or low impedance. In addition, for simplicity of construction, it is desirable to avoid shunt elements such as stub lines, and to use series elements only.

For predictable operation the output spectrum should be essentially a single harmonic. Single diode multipliers offer no way to discriminate between harmonics except through filter structures and are limited by overlapping harmonics in band. For doublers this still permits a very wide band approaching 1.5:1 or 35% fractional bandwidth, but for triplers the fourth harmonic overlap into the band sets a limit of only 28% bandwidth. Practical mm-wave tripler circuits closely approach this limit. Multiple diode circuits exploiting symmetry, called balanced multipliers, have inherent rejection of alternate harmonics. In principle, circuits having a higher degree of balance are possible, in which only harmonics of multiples of 3 or more are produced, but none seem to be practical at mm-wave frequencies.

Balanced doublers are practical up to 400 GHz because their circuits are so easy to realize using a variety of full wave rectifier designs. Balanced doublers require no filters to separate frequencies and some are easy to bias. Balanced triplers, by contrast, require much more complex circuitry which includes filters and difficult bias circuitry. Thus far no successful mm-wave varactor balanced tripler has been reported (with performance even comparable to single diode designs). Using whisker contacted diodes, balanced doublers have a bandwidth limit of 25% due to the limitations of circuit elements available. At this bandwidth, the efficiency is about half that of the fully optimized (narrow band) value. Planar diode doublers appear to be able to do much better because of the integrated approach to the fabrication of the diodes. Bandwidths of 30-35% seem possible based on circuit simulations.

All wideband fixed tuned circuits require very accurate circuit fabrication. Thus far this has been the primary restriction on their use above 300 GHz, and has led to variable performance even in this frequency range.

Balanced doubler design

The basic design of the balanced doubler has been previously described [1], but in this work the design is extended to cover a wide band. This circuit is nearly ideally suited for wide band use because: it requires no filters which can compromise the matching circuits; the input and output circuits are nearly completely decoupled in the design; and the simplicity of the circuit leads to very high inherent efficiency. The whiskered design for a balanced doubler may use two diodes or two contacts to a single diode chip. In either case the diodes are in series across a waveguide in the input circuit. The output circuit consists of a TEM line (coax or stripline) to which the diodes are coupled in parallel. This line acts as a mode filter and is coupled back into a second waveguide for a lower loss connection to the output. The design of a practical balanced doubler is shown in Fig. 2.

The design process for this circuit is mostly cut and try. There is no easy way to model this circuit except with scale models or finite element simulators. Scale modeling is very difficult because two diode ports are needed having the correct phase of excitation. Finite element simulators (such as HP-HFSS) make this job much easier because it is possible to exploit the circuit symmetry and to model only a single diode port. The symmetry plane is an electric wall at

the input and a magnetic wall at the output. This symmetry plane automatically imposes the correct excitation conditions upon the two diode ports.

A troublesome feature of this circuit is a tendency for the second harmonic to excite the TM₁₁ mode in the input waveguide. This mode looks very much like the output coaxial mode but does not require an inner conductor, and so is readily excited. This sets a general constraint on the design of the input waveguide which is to ensure that this mode is below cut-off for a sufficient distance to prevent any loss of power out the input.

The varactor diode is best modeled as a linear element in the circuit design. While it is certainly possible to optimize a circuit of this type using a nonlinear simulator, it is a very inefficient process, and is unlikely to produce a truly optimized result. This is because of the long computing time required and the numerical noise introduced in nonlinear simulators. In addition, a nonlinear simulator is not easily used to visualize the matching problems involved, and so the user can not usefully interact in the optimization process, changing the form of the circuit as required. The linear model for a diode is derived using a nonlinear simulator in a very basic embedding circuit. The process is to derive the optimum embedding impedances at the input and the output for the desired input power level. In this process the varactor bias voltage must be constrained to the low end of the optimum range. If the bias voltage is allowed to freely float, the result may be a rather high Q load. Constraining the voltage may cause a 10-20% decrease in optimized efficiency, but is generally a good trade-off for a reduced Q and increased bandwidth. In fact, even this process is not really needed. A general rule of thumb that serves very well is to model the varactor as a simple series RC circuit at the input and output. At both frequencies assume the capacitance is $0.40 C_j(0)$. This is true only for a varactor pumped with a ~ 15 V p-p waveform, but the value will increase only up to $0.5 C_j(0)$ for much lower drive levels. The choice of resistance is not so simple but is also much less important. This is because the exact resistance value chosen has almost no affect upon the circuit optimization, as long as the value is not too large. The reason is that it is impossible to match to the very low resistance values typical of varactor loads over any significant bandwidth [2], and so achievable circuits nearly always have resistive components higher than that of the varactor. The more carefully the circuit is optimized, the more true this will be. It is generally safe to pick a resistance which is about 0.2 times the varactor reactance at the input, and to then use this as the varactor load at both frequencies. If a very wide band is simulated, it may be helpful to add a $1/f$ frequency dependence to the resistance.

The circuit design process begins by picking a basic geometry for the diode, whisker, output coax and waveguide. This geometry is modeled in HFSS, and solved for the port S parameters in the input frequency band with a waveguide port, a backwall port and a diode port and with an electric symmetry wall. Next solve the same geometry for the output band without the waveguide port, and substituting a coaxial port for the backwall and a magnetic wall for the symmetry plane. This core circuit is shown in Fig. 1. Now terminate the input waveguide port in its own characteristic impedance and do the same for the coaxial port. Use the linear diode model as a load on the varactor port and compute the wideband match at each frequency band, with diode C_j , input backwall location, and center frequency (f_c) as optimization variables. The variation in center frequency is equivalent to an overall scale factor change in the circuit, and saves the additional simulation of scaled structures. Then vary individual dimensions in the circuit structure to see their effect upon the wideband match. Most of these dimensions are too constrained (or are equivalent to linear circuit variables) to produce much difference in results.

The most useful variables are the waveguide width and the coaxial pin diameter. The decoupling of the input and output circuits means that the pin diameter has little effect on the input, while the waveguide width has almost no effect on the output. It is desirable to tune the input circuit so that the waveguide backwall is fairly close to the diode in order to move the unconstrained part of the output matching circuit as close as possible to the source. The pin may have a step in impedance within the waveguide, with the impedance generally lower for a short distance close to the diode. This low impedance section adds capacitance to the output and helps to tune out the whisker inductance. The whisker length is chosen as a compromise between the input and output requirements, with the backwall location and the coaxial pin compensating for the very different reactance at the two frequencies. The optimum geometry chosen should be based upon the inherent wideband match which may be achieved without the aid of additional circuitry, although this serves only as a rough selection criteria.

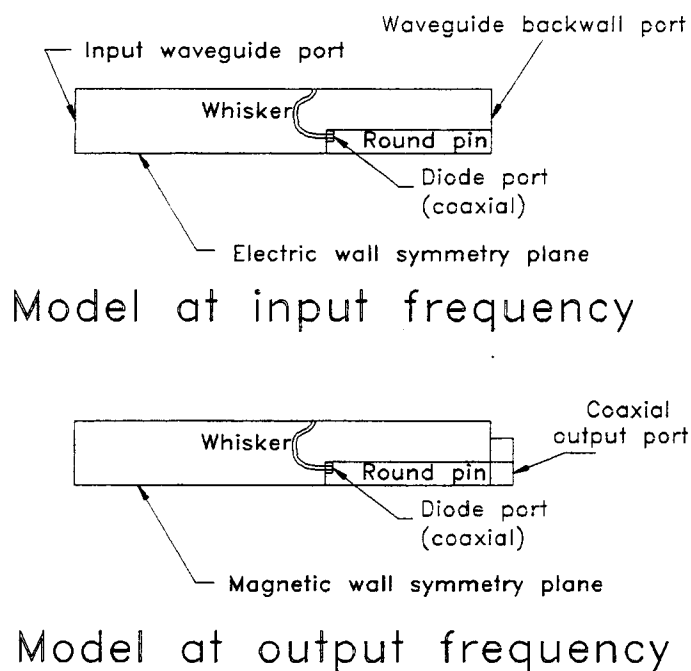


Figure 1. Cross section of the "core" circuit of a balanced doubler. Only half of the waveguide and pin and one diode need to be modeled through the use of symmetry planes.

Having arrived at an optimum "core" circuit model, it is time to design the input and output wideband matching circuits. This process is still iterative, but is much more predictable than the design of the "core" circuit. This still requires as many variables as possible including C_j and f_c , and the final circuit may require a rescaling to match the desired output frequency. In the process of adjusting C_j , keep in mind that the value of C_j strongly affects the optimum operating power level and so is not a completely free variable. However, variations in the whisker length can partially compensate for changing varactor capacitance, so capacitance is a useful variable.

Finally, having arrived at a plausible design, it is necessary to determine if the dimensions chosen can actually be machined using conventional tools. This can be a very extreme constraint on the design if it is to be machined at a reasonable cost, but is not a serious limitation if electroformed construction is allowed. Before actually cutting metal, one might wonder if any

further test is needed to prove the design. Experience with the HFSS simulator has shown that results are sufficiently accurate that circuits will work nearly exactly as designed, and any variations observed are most likely due to errors in fabrication. It may be useful to test the assumptions used in the linear diode model by modeling the full circuit in a nonlinear simulator as a check of the actual performance to be expected. However, none of the circuits to be described have been subjected to any detailed nonlinear simulation. This step is not really essential, and the linear diode model suffices for most purposes.

200-240 GHz Balanced Doubler

The first device to be described is a doubler designed for use as a driver for a tripler for 600-720 GHz. This doubler was intended to produce 10 mW output power, since the tripler was being designed at the same time and the required drive was unknown. The diode chosen was UVa type 6P4 with $C_j(0)=20$ fF, and $R_s=10$ ohms. The optimum input power for this device in this frequency range is 50 mW, and the expected efficiency was about 20%, including the wideband circuit mismatch losses.

The circuit was designed using the procedure described above, and the predicted performance showed a good input/output match over the full 200-240 GHz band, with a VSWR $<2.5:1$. The output broadband match was achieved with a series half wave low impedance line, which has a similar (although lesser) wideband matching effect to a stub line. This half wave line is actually optimized as two quarter λ lines of somewhat different impedance, but it is important to realize that these two line sections do not constitute an impedance transformer. These lines use a round inner conductor, but have a square outer conductor to make the machining easier.

The entire circuit was machined in split block fashion. The layout of the doubler is shown in Fig. 2. Both of the waveguides are split near the E plane centerline, with most of the coaxial outer conductor machined in one half of the block. The whisker pins enter at right angles to the diode. The pins are secured into close fitting slots by slightly peining the edges of the slots. In this improbable geometry, contact of the whisker to the diode is made by setting the whisker slightly off center on the pin and just twisting the pin. This contact may be made without the use of any fixture other than a pair of pliers and a microscope. After the first diode is contacted, the contact to the second anode is determined by the shift in the IV curve and the reduction in R_s .

The diode and the output matching filter pin are supported on a metallized alumina wafer (for insulation) with the pin captured in a channel in a block soldered to the alumina. The use of a slot permits the precise alignment of the pin axially. All of the features in the entire block are machined in the same set-up using a high precision CNC mill. This ensures that there are no alignment errors as can occur if multiple set-ups are used. There are no difficult features to machine except for the output waveguide which has .10 mm width and a depth of .55 mm. This slot is broached with a single tooth cutter in repeated passes under numerical control. The transformer matching to full height WR4 waveguide is machined with conventional mill cutters.

The diodes are contacted with the blocks opened in order to observe the contacting process (which is essential given the unconventional geometry). If all the parts are positioned

properly and the blocks are sufficiently flat, closing up the blocks will not disturb the contacts to typical large area varactors. This is not necessarily true for diodes $< 2 \mu\text{m}$ diameter where the contacts cannot stand small motions.

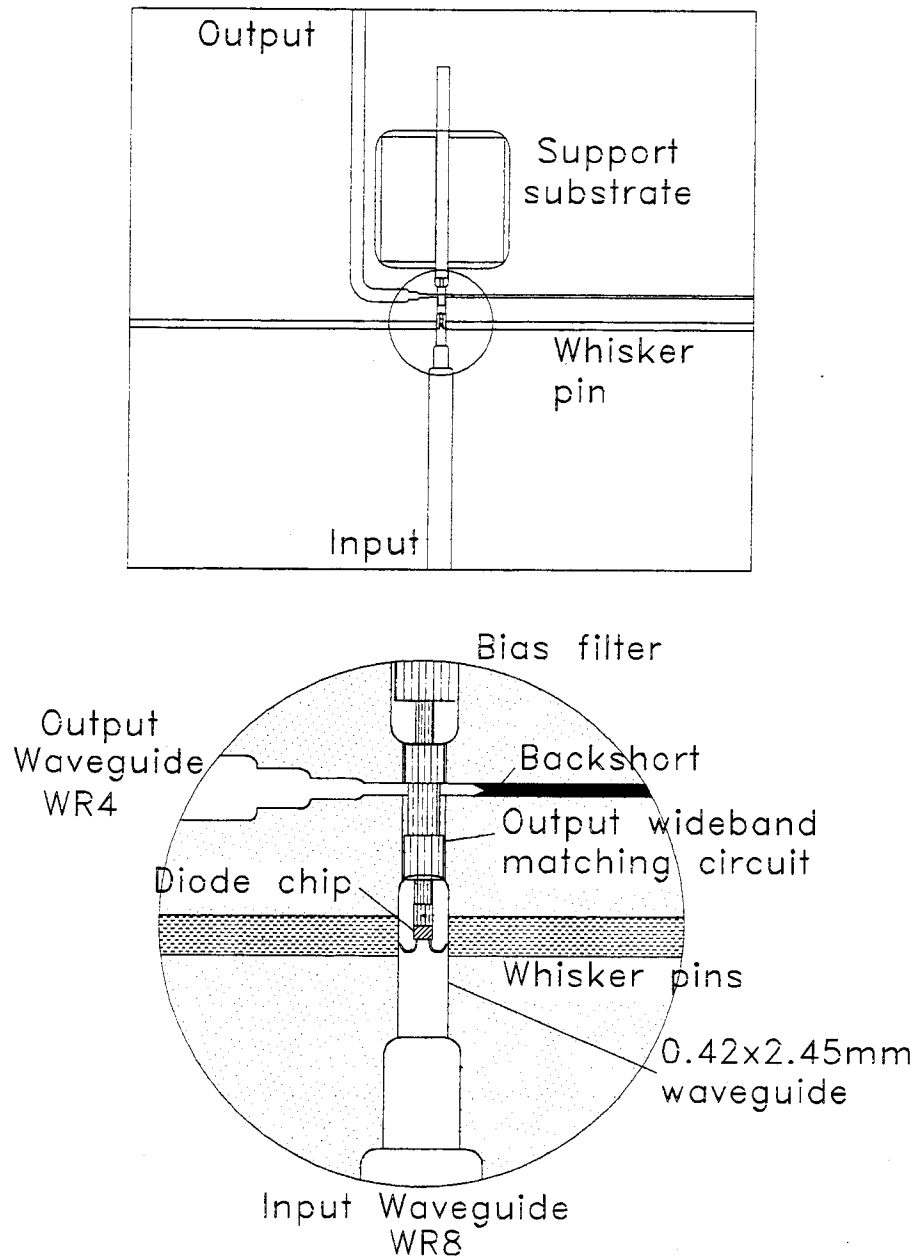


Figure 2. Balanced doubler for 200-240 GHz, with detail of critical area.

There is only one backshort to adjust for final tuning. The best full-band setting is with the short set at the closest position to the coaxial pin which optimizes the output for the highest frequency in the band. There is no other tuning required except to set the bias voltage.

The operational bandwidth exceeds the design band; a range of 194-250 GHz has been tested, but the efficiency is poorly calibrated because no isolator was used on the input. The primary difficulty with measurements is a need for fairly high input power. For an input power below 30 mW, almost no output is obtained while the best results are with >50 mW input. The available Gunn oscillators produced a marginal amount of power and could not produce this much power though the loss of an isolator. It is not correct to assume that the input power is just that which the Gunn will deliver into a matched load since the oscillators are not matched, and may deliver considerably more into the correct mismatch. However, at ~70 mW input near the peak Gunn output, the efficiency is ~20% (midband).

This high power operation arises because the optimum bias for the varactor is 6-7 V, even at low power. For efficient operation the varactor voltage waveform should swing between a small positive (forward) voltage and a large reverse voltage. At 6 V bias this will not happen until around 50 mW input. It is not clear if this situation could have been predicted through a complete nonlinear simulation or if it arises because of an error in assembly (such as excessive whisker length). In any case a lower capacitance varactor ($C_j(0)=16$ fF) would work well at lower power, since it would match at a lower bias, and it should have comparable peak efficiency.

600-720 GHz Fixed-Tuned Tripler

A similar design procedure was used to develop a wide band tripler, which was intended to produce an output in the 600-720 GHz range. This tripler uses a single diode and is similar to others previously described [1,3]. The input matching circuit is based on a coaxial resonator with $\lambda/2$ length at the output frequency. The wideband matching to the input is accomplished with a special coaxial filter design, while the output match with the diode mounted in the output waveguide is inherently wideband. A varactor was chosen with the lowest available capacitance. This device is UVa type 2T8 with $C_j(0)=4$ fF and $R_s=15$ ohms.

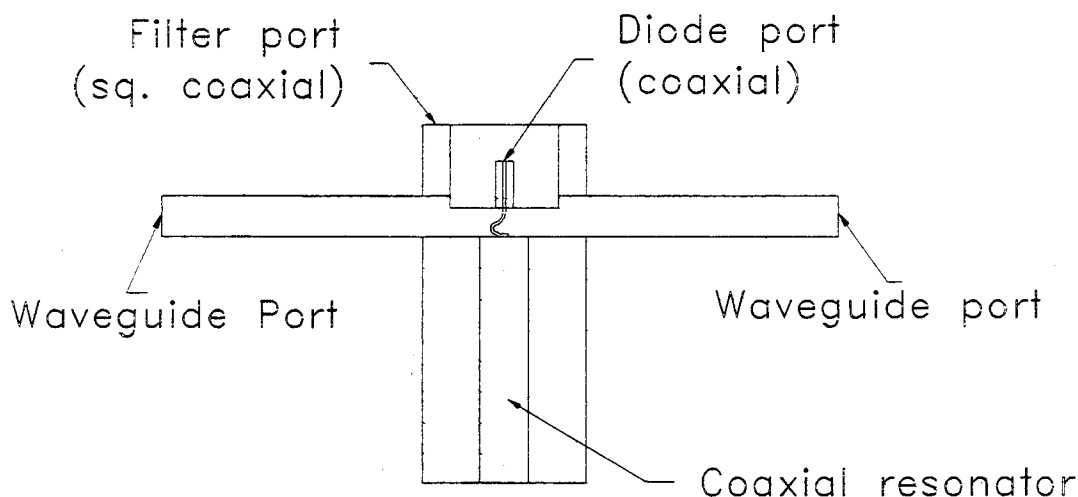


Figure 3. Cross section of "core" circuit of tripler.

The core of the circuit (subjected to HFSS simulation) is the output waveguide where the diode is mounted. This part of the circuit as it would be drawn in HFSS is shown in Fig. 3. In this design a linear diode model was used as in the doubler, with similar resistance and capacitance values. The primary additional circuit requirement for a tripler is a need for an inductive termination at the second harmonic. The output matching was found to be the biggest constraint on circuit design, introduced in part by the relatively high capacitance of the diode. A waveguide height of $<38 \mu\text{m}$ was found to be essential for fixed backshort tuning. In addition an outer diameter to the coaxial resonator of more than 0.15 mm caused too much dispersion in output matching because of the large opening in the waveguide wall.

The coaxial resonator is a good solution to the need for input inductance, but to minimize its effect at the higher harmonics its impedance should be chosen as low as possible, and for this device the impedance is 70Ω . With $\lambda/2$ length there is very little effect in the output band, but the stub is very capacitive in the second harmonic band. This leads to a problem since the optimum idler termination is inductive. Inductance may be added selectively to the second harmonic by enhancing the inductance of the whisker. The whisker coupling to the waveguide behaves as a radial mode resonator for frequencies below cutoff. Increasing the width of the output waveguide near the diode increases the inductance up to the point of resonance at $\lambda/2$ width. In order to suppress the radiation of the second harmonic and to bound this resonator, it is necessary to reduce the waveguide width a short distance away. Including the location of the backshort, a roughly square resonator is formed. While this solution adds inductance, it really only works well near the top of the band, and for lower frequencies there is not enough circuit inductance for best efficiency.

After developing this core circuit, the design of an input matching coaxial line is fairly easy. A five section matching filter was designed using impedances of 14 and 70Ω , with 18Ω along the diode. This filter must have a good (short circuit) stopband at the second and third harmonics but is permitted to have any behavior between the specified bands. The filter is built using a square outer conductor with circular inner sections except for the first section comprising just the diode chip which is square.

The final circuit design was realized in a rather complex block using three pieces which meet at a T joint. The layout of the critical part of the circuit is shown in Fig.4. The remainder of the block is configured much as in the doubler of Fig. 2. The output waveguide is milled as a shallow channel (split along the H plane) in one piece in order to achieve the very low height required. The input waveguide and the coaxial filter were machined in a pair of blocks split along the waveguide E plane for best accuracy. The coaxial filter was supported as in the doubler above. The two input parts must have a very smooth joint where they cross the output waveguide. The coaxial resonator is made using a pin with a shoulder to ensure a good ground. Poor grounding of this pin (even a 1-2 μm gap) can greatly increase the circuit loss. These considerations make the block a very difficult part to machine, but it probably can be simplified as lower capacitance diodes become available or better methods of cutting waveguide are developed.

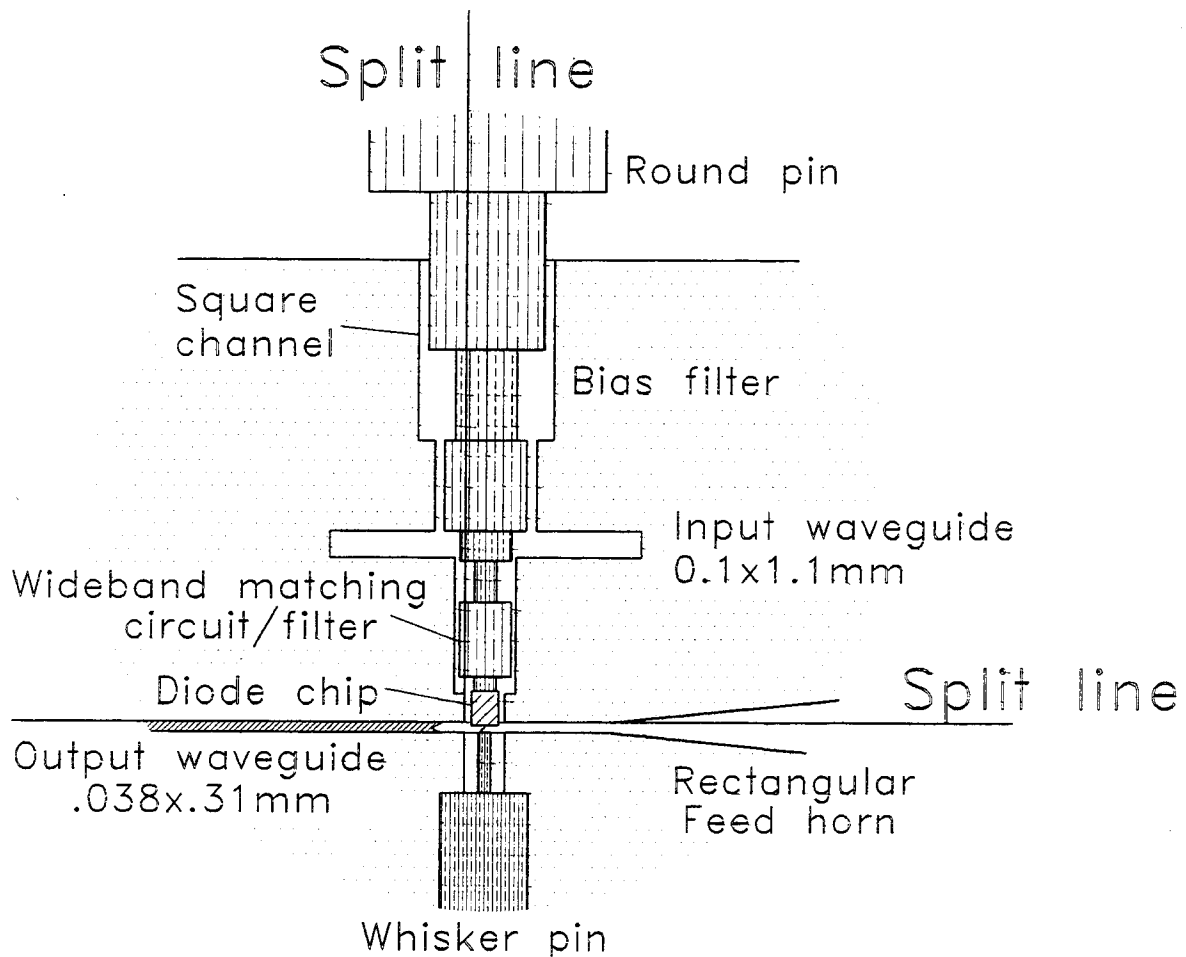


Figure 4. Cross section detail through fixed tuned tripler for 600-720 GHz.

The assembly and contacting process were relatively easy considering the frequency involved. There were no alignment problems between the coaxial pin and the outside, and the contacting was readily done by sliding the choke pin in its groove. Full band operation was obtained without any circuit modification at an efficiency of 0.5-1.0%. Operation is in a varactor mode with a bias of 1.5-3 V at an input power of 3-10 mW. The input and output power are shown in Fig. 5. The individual powers were measured at different times and may not be at exactly the same frequency. The overall shape to the curves is largely due to the available Gunn oscillator power. Power at both frequencies was measured with a calorimeter [3] and should be accurate to within 10%. There is one apparent hole in the output spectrum near 696 GHz, but the input power is also low here. It is not clear if the doubler or tripler is responsible for this hole, or if perhaps some other tuning of the backshorts would eliminate this problem. It is relatively easy to find a set of backshort positions that work but it is difficult to be certain that they are really the best over the full band. In addition, the interactions between the doubler and tripler make it difficult to tune at all frequencies, and require an adjustable phase shifter to obtain full frequency coverage.

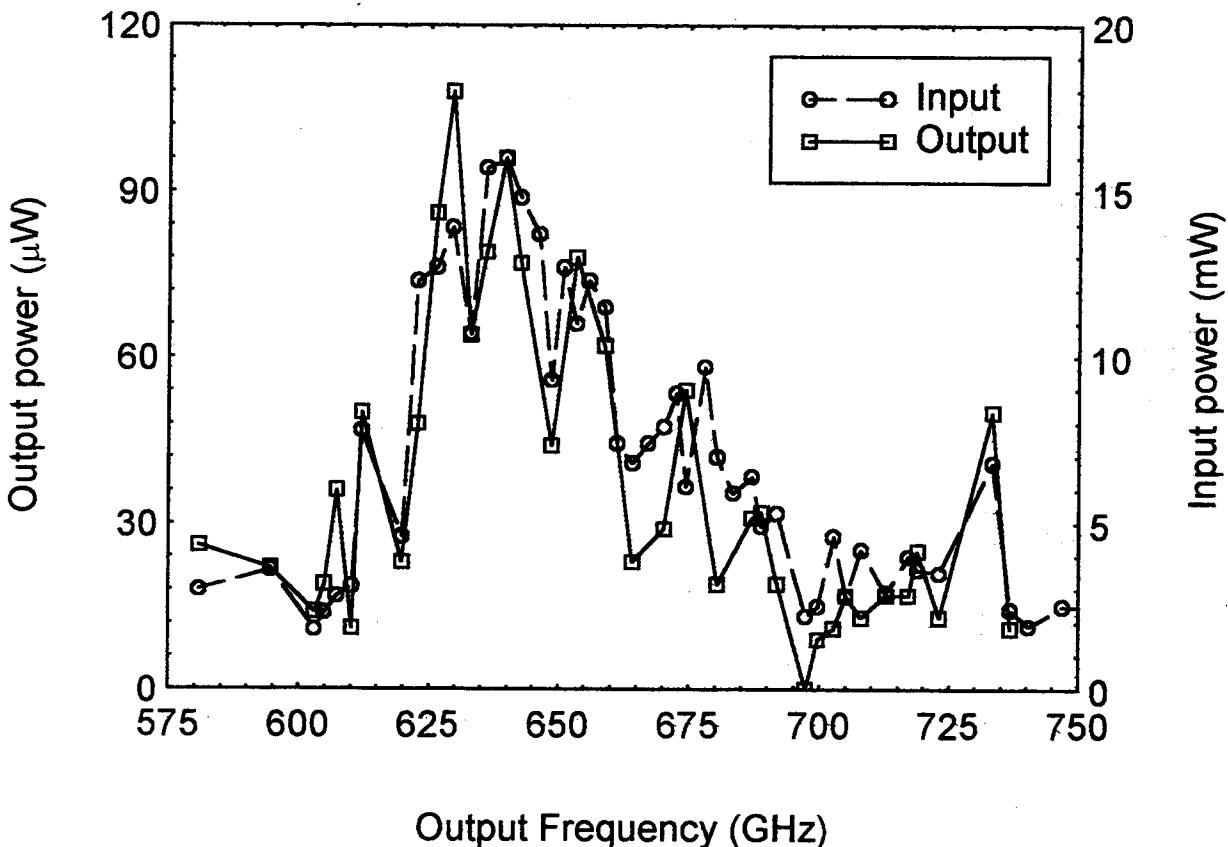


Figure 5. Input and output power for 600-720 GHz tripler. Input power is from 200-240 GHz doubler.

250-320 GHz Fixed Tuned Tripler

A similar design was used in a tripler for 250-320 GHz, but due to the lower frequency the fabrication was much simpler. All of the features in this device were machined in a block split in the E plane with no really difficult features. The output waveguide was broached with a single tooth cutter while the remainder of the output waveguide was milled as full height WR3. The transition between these waveguides included the high pass filter required to suppress second harmonic radiation and was milled as a "channel transformer" with varying depth milled features in a cross section resembling a + sign [4]. The channel transformer is very useful device for this application because the waveguide cutoff frequency in the channel region is always higher than that of the uniform rectangular cross section at the ends. This cutoff frequency can be varied using just the depth of one cut relative to the other. However the cutoff frequency and the impedance are tied together so it is a bit tricky to design an impedance transformer with a given impedance ratio and cutoff frequency.

The assembly was very easy relative to previous wafer style triplers at a similar frequency and the initial performance matched expectations. The input backshort was easily set using a measurement of the small signal input return loss. This small signal tuning works well because the

series resistance of the diode is typically only a factor of three smaller than the optimum input resistance under large signal drive. As long as the bias voltage is set to produce the correct effective capacitance (typically 4-5 V), the general shape of the frequency response of S11 is essentially independent of power level except that its magnitude will decrease at higher power. The output backshort is tuned for the best power output at the maximum frequency in the band, as this has typically been found to optimize the full band. The overall performance is excellent, and is comparable to older mechanically tuned models optimized at each frequency over a similar band.

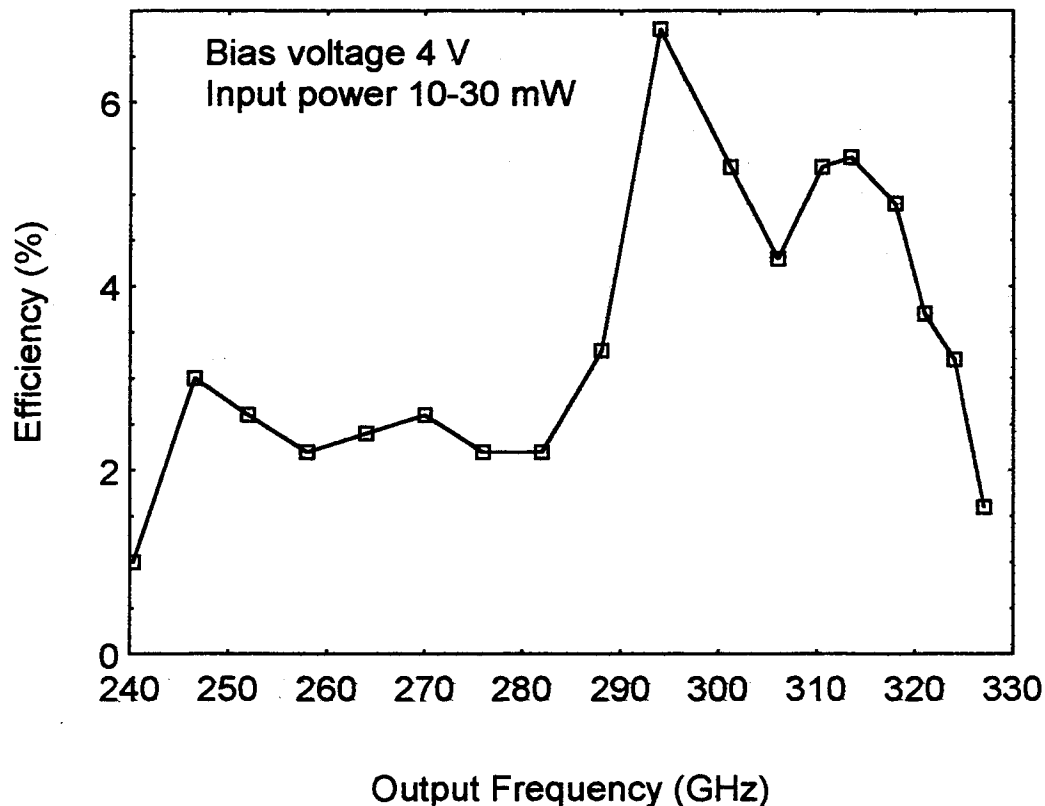


Figure 6. Efficiency of fixed tuned tripler for 250-320 GHz.

Cascading multipliers and interstage interactions

Cascading of multipliers is difficult because all wideband varactor multipliers are inherently poorly matched. An input/output VSWR = 2.5-3:1 is typical for the devices described above and this will degrade further at low power levels or if the bias is set incorrectly. Cascading of circuit elements with a VSWR >2:1 produces significant power ripple, and generally is undesirable even for linear elements. Varactor devices present additional problems because of their nonlinearity, and are even prone to instabilities such as parametric oscillations [5]. Typical cascaded multipliers measured over a wide band will show some frequencies where useful output may not be obtained or where power is very sensitive to frequency. Adding a tunable phase shifter between stages generally permits operation at any frequency, since it eliminates the worst case interaction, but it adds loss and requires mechanical tuning.

Ideally one would add an isolator between all stages, but mm-wave wideband isolators have fairly high loss, and are not available at all above 220 GHz. While a loss of 2 dB may be marginally acceptable at 100 GHz, a loss of 4.5 dB is not at 200 GHz, and so isolators are not really a good option at present.

In principle the input and output may be matched through the use of a balanced construction using wideband 90° hybrids on the input and output with parallel multiplier circuits. This only works well if the multipliers are identical in efficiency and phase, and this may never be true at high frequencies. In addition, the bandwidth of the hybrids is limited and the required operating power doubles.

This unsatisfactory situation is now improving. In recent work the author has made wideband isolators with much lower loss than previous models, although they use the same basic design as present Faraday rotation isolators [6]. The differences are a much shorter ferrite length (with a correspondingly higher magnetic field), and much reduced lengths of waveguide on either end. The loss of a WR10 model is <1.0 dB across the band, while a WR6 model has typically <1.5 dB midband loss with 2 dB worst case, and this loss can probably be reduced further. It even seems possible to build a WR3 model, although the loss may be significantly higher.

Conclusions

These devices represent a sample of designs that are now possible for wideband applications. These same designs can in principle be scaled to any frequency but peculiarities of each frequency range usually require some redesign. The balanced doubler with whisker contacts is feasible up through at least 400 GHz, while a planar diode version has been designed with very wide bandwidth at 150 GHz.

The tripler to 720 GHz is pushing the limits of the techniques used but is practical at any lower frequency. Lower capacitance varactors may extend the frequency upward, while with reduced bandwidth its fabrication becomes somewhat easier.

These types of devices should be able to completely replace tunable multipliers in the near future. With the increasing capabilities of planar diodes and new designs using them, many of these same circuits should soon become feasible with planar diodes up through 400 GHz.

References

- [1] Erickson, N.R., "High efficiency submillimeter frequency multipliers," 1990 IEEE MTT-S Int'l. Microwave Symposium, p. 1301.
- [2] Matthaei, G.L., Young L., and Jones E.M.T., "*Microwave Filters, Impedance-Matching Networks and Coupling Structures*," (section 4.09), Artech House Books, Dedham, MA, 1980.
- [3] Erickson, N.R., and Tuovinen, J., "A waveguide tripler for 720-880 GHz," Proceedings of the Sixth Int'l. Symposium on Space THz. Tech., pp. 191-198, Mar. 1995.
- [4] Siegel, P.H., Peterson, D.W. and Kerr, A.R., "Design and analysis of the channel waveguide transformer," IEEE Trans. Microwave Theory Tech., June 1983, pp. 473-484.
- [5] Penfield, P. and Rafuse, P.R., *Varactor Applications*, MIT Press, Cambridge, MA, 1962.
- [6] Millitech Corporation, S. Deerfield, MA series FBI isolators.