A Novel Quasi-Optical Frequency Multiplier Design for Millimeter and Submillimeter Wavelengths

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Abstract —This paper describes a novel design for millimeter and submillimeter wavelength varactor frequency triplers and quadruplers. The varactor diode is coupled to the pump source via waveguide and stripline impedance matching and filtering structures. Output power at the various harmonics of the pump frequency is fed to quasi-optical filtering and tuning elements. The low-loss quasi-optical structures enable near-optimum control of the impedances seen by the varactor diode at the idler and output frequencies, resulting in efficient high-order harmonic conversion. A minimum efficiency of 4 percent with 30-mW input power has been obtained for a tripler operating between 200 and 280 GHz, with a peak efficiency of 8 percent between 250 and 280 GHz. Another tripler, designed for the 260–350-GHz band, gave a minimum conversion efficiency of 3 percent with 30-mW input power, with a peak efficiency of 5 percent at 340 GHz.

I. INTRODUCTION

FREQUENCY MULTIPLIERS are currently widely used to provide a reliable, and relatively inexpensive, source of local oscillator power in millimeter and submillimeter wavelength heterodyne receivers [1]–[3]. In the past few years, improvements in varactor diode characteristics and diode mount design have resulted in significant improvement in multiplier efficiency and output power [4]–[6]. Most of the multiplier mounts reported have utilized waveguide, stripline, or coaxial structures for tuning and filtering at the input, output, and idler frequencies. This approach has proven useful for frequency doublers, where no idler is required, at output frequencies up to 700 GHz [7]. However, in the design of higher order harmonic generators to operate at output frequencies above 200 GHz, the waveguide and stripline structures required to optically tune the various harmonic frequencies for maximum conversion efficiency become extremely difficult to fabricate, especially if it is required that the multiplier be tuneable over a wide frequency range. This problem is overcome in the device described here by using quasi-optical elements for idler and output frequency tuning and filtering. This approach is eminently suited to LO applications in millimeter and submillimeter systems in which quasi-optical diplexing structures are commonly employed [3].

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The paper commences with a description of the waveguide and stripline structures which bring the power to the diode at the pump frequency. The practical constraints on this aspect of the design are discussed using the tripler mount for 200–280-GHz output, as an example. The higher frequency mount, for 260–350 GHz, is a geometrically scaled version ($\times 0.80$) of the lower frequency design. The varactor diode, mounted in a broad-band waveguide structure, is coupled to the quasi-optics via a feed horn and dielectric collimating lens. The design and performance of the quasi-optics in the output circuits is considered and, finally, RF performance data is given for several complete frequency tripler and quadrupler assemblies operating in the 200–350-GHz output frequency range.

II. WAVEGUIDE AND STRIPLINE STRUCTURES

Efficient coupling between the pump source and the varactor diode at the input frequency is an essential factor in the minimization of the harmonic conversion loss of any frequency multiplier [8]. Furthermore, power at the harmonic frequencies should be isolated from the pump circuit [8]. For pump frequencies in the range 65 to 120 GHz, these functions are most readily carried out by waveguide and stripline structures.

The varactor diode mount, shown schematically in Fig. 1, employs a crossed-waveguide, split-block construction similar in some respects to previously reported designs [1], [9]. Pump power incident in the full height input waveguide is fed via a doubly tuned transition to a stripline low-pass filter, which is designed so that it passes the pump frequency with low loss, but presents a high attenuation to higher order harmonics. The whisker-contacted varactor diode is attached to the output end of the low-pass filter substrate, adjacent to the output waveguide. Pump circuit impedance matching is achieved using two adjustable waveguide stubs with sliding contacting shorts. One stub acts as a backshort for the probe-type waveguide to stripline transition and the second as an E-plane series stub, located approximately $\lambda_g/2$, at the pump center frequency, towards the source from the plane of the transition. This tuning configuration allows pump circuit matching with a VSWR of typically 2.0:1 or less at any frequency in the design band (67 to 93 GHz in WR12 waveguide).

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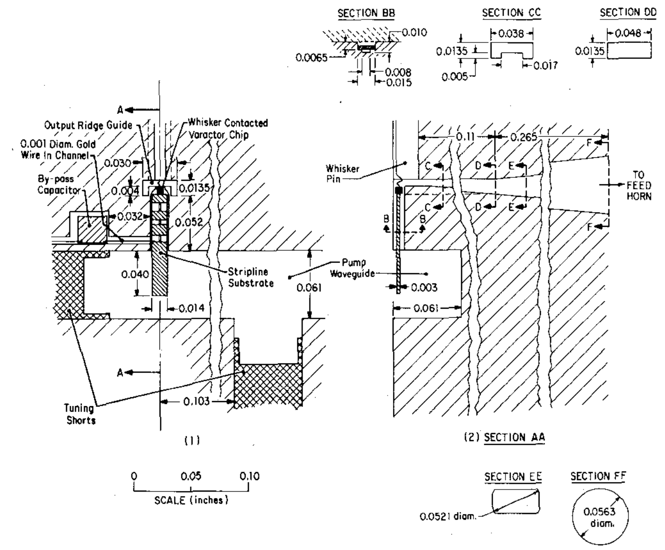

(a) (b)

Fig. 1. (a) A view of the face of the multiplier block containing the diode. The upper block forming the complete mount has been removed for clarity. The drawing is approximately to the scale indicated, with dimensions being in inches. (b) Cross-sectional views through various parts of the block shown in Fig. 1(a). The waveguide transformer dimensions and stripline channel geometry are detailed.

The stripline low-pass filter is a seven section, 0.2-dB ripple Chebycheff design, implemented using high/low impedance stripline sections printed on a crystalline quartz substrate. The cutoff frequency is 110 GHz. Performance and design data have been given elsewhere [9]. The varactor diode chip, cut to a 0.005-in-sided cube, is soldered to a gold tab attached to the low-pass filter metallization. The last capacitive section of the choke is shortened to compensate for the capacitance between the chip and the sides of the stripline channel, since the chip is mounted in the channel with its face flush with the output waveguide broadwall. The 0.0005-in diam, phosphor bronze contact whisker, suitably pointed and bent, is mounted on a 0.02-in diam beryllium copper post, which is an interference fit in a hole bored through the block to the broadwall opposite the diode chip. The whisker is bent so that it just spans the guide and the pin sits flush with the waveguide wall after contacting the diode. The unbent length of the whisker is empirically optimized for best performance and depends upon the characteristics of the varactor diode employed and upon the output frequency band. It is typically 0.007 in for the mount designed for the 200–280-GHz output band.

The varactor diodes used in this work were 6-μm diam Schottky-barrier devices supplied by R. J. Mattauch of the University of Virginia. Designated type 6P2, they have a typical zero-bias capacitance of 20 fF, a dc series resistance of 8 Ω, and a reverse breakdown voltage of 18 V at 1 μA. The vapor phase epitaxial material upon which the diodes were fabricated has a nominal doping of 1.6×10^{16} cm^{-3} and an epi-thickness of 1.1 μm. These devices have a highly nonlinear capacitance versus voltage law which approximates the inverse half-power behavior of the ideal abrupt junction varactor to within about 2 V of the breakdown limit.

The output circuit of the multiplier must provide efficient coupling between the low dynamic impedance of the pumped varactor diode (real part of the order of 25 Ω) and a load at any frequency in the desired output band. In addition, in a frequency tripler, the output circuit elements should reactively terminate the varactor at the second harmonic to prevent power being coupled to the load at the idler frequency [8]. In the present design, these tuning functions are performed quasi-optically and it is sufficient, therefore, for the output waveguide in the vicinity of the diode to have a low wave impedance and a broad single-mode bandwidth, extending from the lowest second harmonic frequency to at least the highest third harmonic frequency expected. Such a structure maximizes the frequency range over which the nonlinear device can be efficiently coupled to the output filtering and matching networks. In the mounts described here, the varactor diode is coupled via the contact whisker to a single ridge waveguide. The guide dimensions for the lower frequency unit are shown in Fig. 1. The single-mode bandwidth of this structure, in the vicinity of the diode, extends from 123 to 292 GHz for the mount designed for the 200 to 280-GHz output band. The wave impedance of the guide varies minimally (from 113 to 99 Ω) over the output frequency range of 200 to 280 GHz [10]. A fixed backshort in the single ridge guide is positioned a distance of about $\lambda_g/4$ (at the output center frequency) from the plane of the diode. The position of the short relative to the diode does not have to be accurately set since the multiplier output is tuned quasi-optically.

Power in the single ridge guide near the diode is coupled out of the mount to the quasi-optical structures via a tapered waveguide transformer and a broad-band scalar feed horn. The waveguide ridge height and width first undergo a linear taper to reduced height rectangular waveguide of dimension 0.048 in $\times$ 0.0135 in. This simply fabricated form of taper maintains a constant TE_{10} mode cutoff wavelength at any cross section of the transition and, for a sufficiently long taper (longer than $\lambda_g/2$ at the lowest operating frequency), results in minimal spurious mode coupling [10]. The reduced height rectangular guide, which can be multimoded above 246 GHz, is then smoothly coupled to the circular waveguide at the throat of the feed by a further linear flaring of the guide dimensions. The output waveguide structures and feed horn were fabricated by electroforming copper on disposable aluminum mandrels.

III. Feed Horn and Lens Design

The clear aperture diameter through the quasi-optical components was set at 2.0 in for compatibility with an existing diplexer design [3]. The choice of this dimension constrains the maximum waist width for low-loss propagation of the electromagnetic beam within the tuner and hence determines the design of the collimating lens and illuminating feed horn. The full width of the beam at the $1/e$ points is usually made equal to half the clear aperture diameter for minimum loss due to diffraction [11].

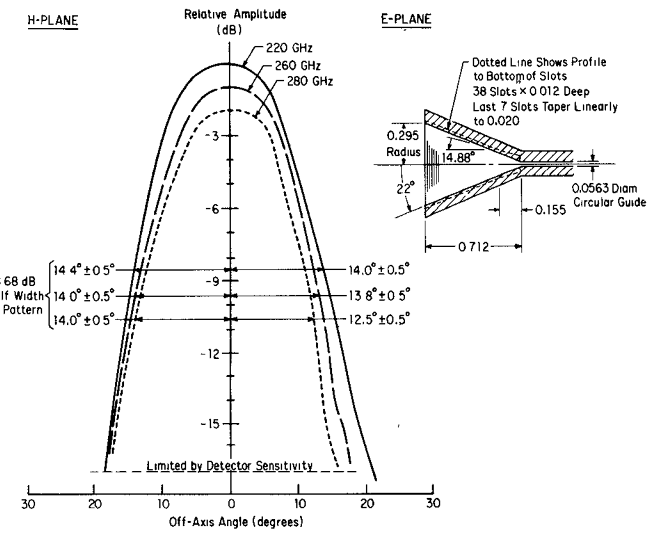

Fig. 2. Experimental measurements, at several different frequencies, of the far-field radiation pattern of the corrugated horn feed used in the multiplier. The horn dimensions are shown schematically in the inset. The H-plane half pattern is shown on the left and the E-plane half pattern on the right. The measured $1/e$ width of the patterns is indicated. The 0-dB reference for each pattern has been arbitrarily offset for clarity.

In the present case, for optimum performance of the quasi-optical tuner and filter, the beam size in the near field of the lens should be approximately constant at this nominal value over the range 134 to 280 GHz. Furthermore, the electromagnetic beam incident on the quasi-optics should have a plane wave front in this region. This requirement implies that the illuminating feed horn should maintain a constant beamwidth and a fixed phase center over the design frequency range; conditions which can be approximately satisfied by a circumferentially corrugated conical horn [12] with a sufficiently large flare angle. If in such a horn, radiating the balanced HE_{11} mode, the peak phase deviation of the spherical wave front from a plane wave over the horn aperture is larger than about 0.75 λ_0, then the beamwidth is essentially frequency independent and is determined largely by the semi-angle of the horn. The phase center is located near the throat. The main lobe of the far-field radiation pattern of the horn possesses circular symmetry and has a radial amplitude distribution which is well described by a simple Gaussian function. In addition, the sidelobe levels are usually more than 30 dB below the peak of the main beam. Hence, this type of horn is ideally suited for use as a primary illuminator in a quasi-optical system employing fundamental mode Gaussian optics [11].

The 0.59-in aperture diameter horn used for the multiplier, shown in Fig. 2, has a semi-angle of 22° and is designed for optimum performance in the 200–280-GHz band. However, the horn characteristics from 130 to 187 GHz result in adequate performance in the second harmonic band as well. The collimated peak aperture phase deviation varies from 0.65 λ_0 at 134 GHz to 1.358 λ_0 at 280 GHz. Assuming radiation from a balanced HE_{11} mode, it is evident that the $1/e$ width of the ideal far-field pattern for this horn (29°) is essentially independent of frequency

TABLE I
THEORETICAL PROPERTIES OF THE CORRUGATED HORN AS A FUNCTION OF OPERATING FREQUENCY

Parameter	Frequency GHz			
	134	180	230	280
$\Delta = \frac{a}{\lambda} \tan\left(\frac{\theta_o}{2}\right)$	0.65	0.87	1.12	1.36
$\frac{2\pi a\theta_o}{\lambda}$	8.074	10.844	13.363	16.879
$\frac{X_{slot}}{Z_o} = \delta \tan\left(\frac{2\pi d}{\lambda}\right)$	0.72	1.45	9.15	-2.24
γ	0.60	0.80	0.93	1.09
E-plane pattern 1/e full width	34.9°	31.9°	30.2°	28.3°

NOTES: H-plane 1/e full-width is essential frequency independent at 29.3°

a = aperture radius   d=slot depth δ=slot width/pitch ratio

θ_o = horn semi-angle γ = mode content factor - ratio of longitudinal fields of TE_{11} and TM_{11} components

Δ = aperture phase deviation, edge to axis

—the main effect of the changing phase deviation is to alter the shape of the response near the axis of symmetry [12].

The balanced HE_{11} mode exists only in a horn in which the longitudinal surface impedance is infinite, i.e., the circumferential slots are near $\lambda/4$ deep. As the frequency deviates from slot resonance, the balance between the TE_{11} and TM_{11} mode amplitudes comprising the HE_{11} mode is disturbed, degrading the beam circularity and increasing the cross-polarized component. The slot resonant frequency for the horn described here was chosen to be 240 GHz, corresponding to a slot depth of 0.0123 in. The curves given by Thomas [13] may be used to predict the deviation from beam circularity and the increase in the cross-polar component as the operating frequency deviates from 240 GHz. Table I shows the results of this analysis, indicating that the maximum difference in E and H plane -8.68-dB beamwidth occurs at the lowest operating frequency (134 GHz) and is about 25 percent of the full width at 240 GHz. The slots in the wall of the horn are fabricated with a period of 0.020 in (close to $\lambda/2$ at 280 GHz) with a wall thickness between slots of 0.008 in. A thinner wall would be desirable for optimum cross-polarization performance [12]. However, practical constraints in the electroforming process make it difficult to fabricate a horn with thinner partitions. In order to improve the horn VSWR and reduce coupling to the EH_{12} mode in the throat region, the depths of the first 10 slots taper linearly from 0.020 in to the final 0.0123 in.

Fig. 2 shows the measured far-field pattern of a practical feed at several different frequencies in the band of interest. The performance agrees quite well with the theoretically predicted behavior.

The feed horn illuminates a 2.00-in diameter circularly symmetric collimating lens made from Teflon. The lens is designed on the basis of geometrical optics and is constructed so that the surface towards the feed is plane. The lens thickness, at a given radial distance from the center, was derived from the parametric formulas given by Silver

[14]. The focal length (1.726 in) of the lens is chosen to provide the correct beam waist width in the quasi-optics when illuminated by the feed described above. The lens surfaces are concentrically grooved in order to reduce reflection losses at the air-dielectric interface. The grooves have an easily machined triangular cross section [15] (included angle 43.6°, depth 0.012 in, pitch 0.015 in) which results in a power reflection coefficient for the lens of less than 0.01 over the 200–280-GHz band.

IV. Quasi-Optical Filtering and Matching Structures

For best conversion efficiency in a high-order varactor frequency multiplier, the intermediate harmonic frequencies (idlers) must be reactively terminated [8]. To enable the multiplier to operate well over a wide range of frequencies, the idler terminations should be readily tunable. In a waveguide structure it is difficult to realize tunable idler terminations which are decoupled from an output frequency tuning circuit. The quasi-optical approach described here, and shown schematically in Fig. 4, overcomes these problems, providing the means for separately tuning the idler and output circuits.

The idler termination is implemented using a dichroic mirror as a high-pass filter. The dichroic plate comprises an aluminum sheet of accurately determined thickness (L), perforated with an equispaced array of holes (spacing S) of precisely machined diameter (D). The device may be represented by an equivalent circuit, shown in Fig. 3, which comprises a length (L) of transmission line (circular waveguide of diameter D) shunted at each end by an inductive admittance. For wavelengths below cutoff in the circular guide, the plate exhibits very low transmittance and behaves as a plane mirror. Above cutoff the power transmission (T) is given, for normal incidence, by [16]

$$T=4\left\{4\left[C-\frac{B_S}{Y_2}S\right]^2+\left[\frac{Y_1}{Y_2}S+2\frac{B_S}{Y_1}C+\frac{Y_2}{Y_1}S-\frac{B_S^2}{Y_1Y_2}S\right]^2\right\}$$

where

$$C=\cos(\beta L),\ S=\sin(\beta L)$$

$$B_S=1.096\times10^{-3}\left(\frac{S}{D}\right)^2\left(\frac{\lambda}{D}\right)\left[1-\left(\frac{1.706D}{\lambda}\right)^2\right]\ \text{ohm}^{-1}$$

$$Y_2=3.496\times10^{-3}\left(\frac{S}{D}\right)^2\left[\frac{1-(0.426D/S)^2}{2J_1'\left(\frac{\pi D}{4S}\right)}\right]\frac{\lambda}{\lambda_g}\ \text{ohm}^{-1}$$

$Y_1=2.652\times10^{-3}\,\text{ohm}^{-1}$ (the free-space admittance)

λ, λ_g = free space and guide wavelength, respectively.

The model shows that for unity transmission, L must be $\lambda_g/2$ if $B_S=0$. Since B_S is nonzero in practice, the plate will be somewhat thinner than $\lambda_g/2$ at the transmission center frequency.

For a successful dichroic plate design, two other factors must be taken into account. Firstly, since the plate behaves in a transmission as a radiating array of equispaced circular waveguides, the spacing between the holes must be chosen to be sufficiently small for the first grating lobe response to lie at an angle of at least 90° to the direction of propagation. This implies the constraint, in an air-dielectric environment [16]

$$\left(\frac{S}{\lambda}\right)_{\max}=\frac{1.1547}{(1+\sin\theta)}.$$

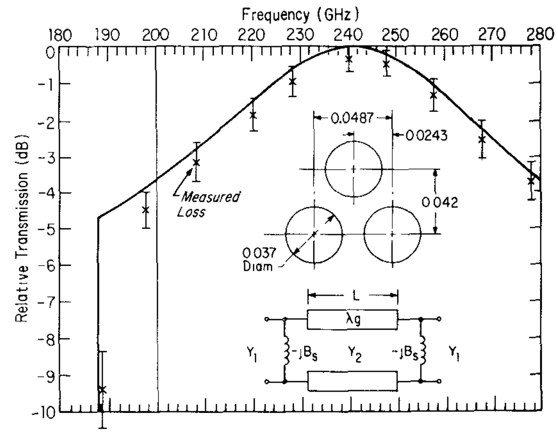

Fig. 3. The theoretical and measured response of the dichroic plate. The significant dimensions (in inches) of the array of holes are shown in the inset. The thickness of the plate was 0.030 in. The equivalent circuit used in the analysis is also shown. The errors in the measured transmission are estimated on the basis of the unavoidable impedance mismatches in the measurement setup

Secondly, the first passband transmission maximum should not be too close to the cutoff frequency of the plate; otherwise, resistive losses in the circular waveguide segments become significant.

The plate used in the present system was designed to have a cutoff frequency of 186.7 GHz and a transmission center frequency of 240 GHz. The hole diameter is 0.037 in positioned in an equilateral array on 0.0487-in centers. The plate thickness is 0.030 in. The theoretical and measured transmission response of the plate is shown in Fig. 3. The plate was made from a piece of 6061 aluminum, precisely milled to the correct thickness, and the holes were drilled using a special rotary fixture.

In order to ensure that it is flat, the dichroic mirror is mounted on a special carrier fabricated from a commercial stainless steel, 32 pitch gear, 2.563 in in diameter. The central hub of the gear is machined away, leaving an angular ring with a 2.00-in bore. A precision 32 T.P.I. thread is cut across the top of the gear teeth. The externally threaded gear is mounted in a housing which has a matching precision thread in an internal bored hole. The arrangement is schematically shown in Fig. 4. The lens, feed horn, and varactor mount are mounted on a special fixture attached to the end of the housing, as shown in Fig. 4. The output signal from the varactor is injected axially by the feed horn and lens along the bore in the housing. The spacing between lens and feed can be readily adjusted by turning a knurled adjusting ring, which has internal opposed left- and right-hand threads which mate with similar threads on the lens holder and block mounting section. Pins prevent the two halves of the fixture from rotating relative to one another as the ring is turned.

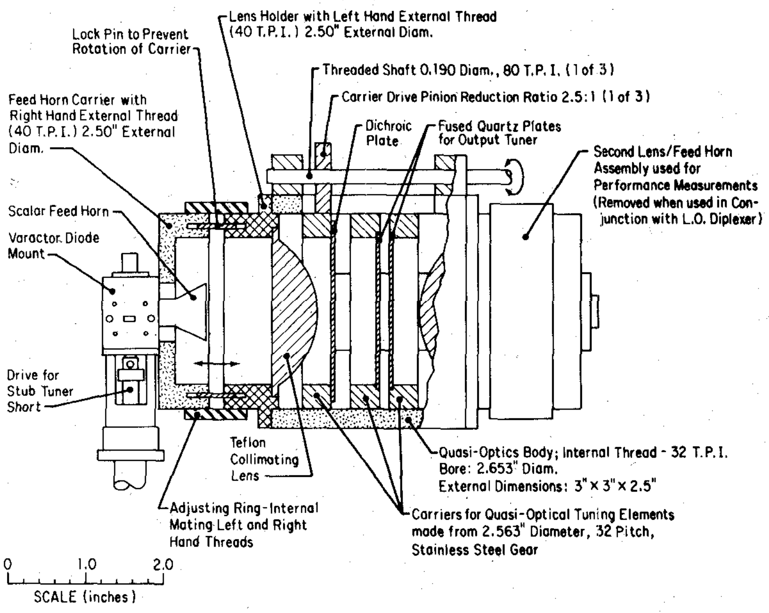

Fig. 4. A schematic diagram of the assembled quasi-optical frequency tripler. The drawing is approximately to the scale indicated.

To tune the idler frequency termination, the dichroic mirror must be moved axially relative to the varactor mount, thus varying the phase of the reflected signal at the plane of the varactor. It is clear that the mirror must be flat and square to within about $\lambda/10$ at the highest operating frequency and that the incident beam must consist of plane wave fronts if the reflected signal is to be efficiently coupled back to the mount. Accurate translation of the plane is achieved in this design by rotating the threaded gear carrier using a secondary drive pinion. This achieves a fine linear motion of the plate, precisely square with the translation axis, at a displacement rate of 0.0125 in per turn of the drive pinion.

Output frequency matching is achieved with the aid of a quasi-optical dual dielectric plate tuner. Power at the output frequency, after passing through the dichroic reflector, is incident on a pair of 0.0189-in thick fused quartz plates. The plates, with surfaces ground flat and parallel to better than $\lambda/20$ at the shortest output wavelength, are mounted on carriers similar to that which hold the dichroic mirror. The tuning plates may thus be positioned as desired relative to the varactor mount and to each other. This type of tuner behaves in a similar fashion to a double stub tuner in a coaxial line or a waveguide. An equivalent circuit may be formulated as shown in Fig. 5 in which the quartz plates are represented by fixed lengths of transmission line of impedance $Z_1 = (\mu_0/\epsilon_r\epsilon_0)^{1/2}$. These transmission lines are interconnected by a variable length line (L_2) of impedance $Z_0 = (\mu_0/\epsilon_0)^{1/2}$ and to the load (Z_L) by a variable-length line (L_1) of the same impedance. Since, at any frequency, L_1 can always be adjusted to make the impedance at the tuner real valued, regardless of the value of Z_L, the analysis can be simplified to an investigation of the matching ability of the tuner when presented with a real valued load impedance.

In the present design, the relative dielectric constant of the plates is 3.8 and the thickness corresponds to $3\lambda/4$ at

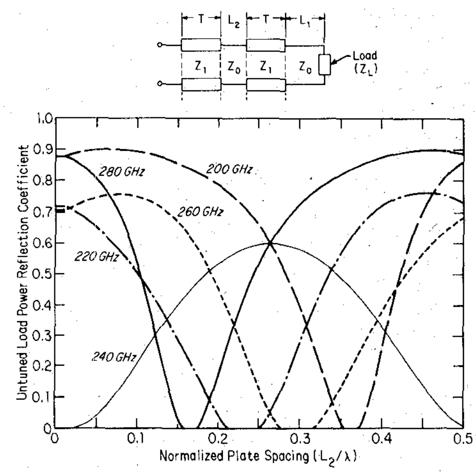

Fig. 5. Expected performance of the quasi-optical dual plate tuner derived from an analysis of the equivalent circuit shown in the inset.

240 GHz. An analysis of the equivalent circuit results in a plot, shown in Fig. 5, of the magnitude of the load power reflection coefficient that can be matched versus spacing between the plates for several different frequencies in the 200–280-GHz range.

V. Performance of Some Complete Multiplier Assemblies

Two frequency triplers have been constructed using the techniques described above, covering the bands 200–280 and 260–350 GHz. In each case, power in the output beam was refocused by an identical Teflon lens assembly into a second feed horn, which was connected via a short length of single-mode rectangular waveguide to a power meter head.[1] The measured output powers quoted here have been corrected for the transmission losses in the second feed horn and lens, determined by halving the loss measured

[1]Anritsu, Inc., Models MP84B1 and MP86B1.

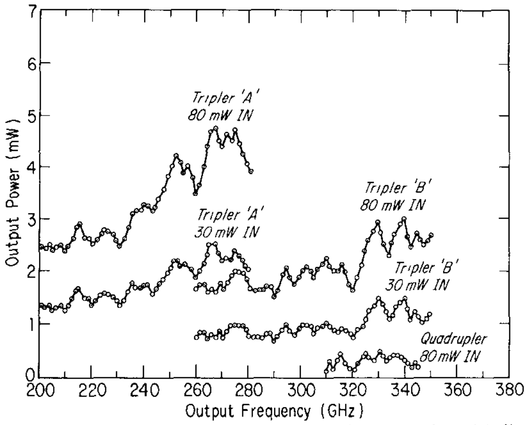

Fig. 6. The measured performance of several practical multiplier assemblies. Tuning and bias were optimized at each measurement frequency.

using two such assemblies back to back. The measured loss varies from 0.3 dB at 200 GHz to 0.6 dB at 350 GHz.

The performance of the multipliers was measured at 1500-MHz intervals for output frequencies between 200–280 and 260–350 GHz, respectively. The output power response as a function of output frequency, for constant 80-mW pump power (the maximum safe pump level), is shown in Fig. 6. The tuning of the input backshorts, the output and idler quasi-optical elements, and the dc bias were adjusted for maximum output power at each measurement frequency. Tuning of the assembly was found to be straightforward, with little interaction between the tuning elements at the various frequencies. Typically, the bias conditions for optimum performance were a reverse voltage of 3.5–4.0 V and a forward current of 0.5 to 1.0 mA.

More than 1.5-mW output power is obtained at any frequency between 200 and 350 GHz. The 200–280-GHz band device provides a minimum output power of 2.4 mW, corresponding to a minimum efficiency of 3 percent for 80-mW pump power. Maximum output power of 4.6 mW is obtained near 265 GHz, corresponding to an efficiency of 5.75 percent. The tripler for the 260–350-GHz band has a minimum output power of 1.5 mW and a maximum output power of 3.0 mW at 340 GHz, for 80-mW pump power. The corresponding conversion efficiencies are 1.875 and 3.75 percent, respectively.

Higher conversion efficiencies may be attained with lower pump powers. The maximum efficiency typically occurs for pump powers of about 30 mW. Fig. 6 also shows the output power responses for a pump level of 30 mW. The maximum efficiencies for the two mounts under this condition are 8 percent near 265 GHz and 5 percent at 340 GHz, respectively.

By installing an additional dichroic plate to reflectively terminate the third harmonic, and using the diode mounting block for the 200–280-GHz band tripler, an experimental frequency quadrupler to the 310–345-GHz band was also constructed. The output power response for this multiplier is shown in Fig. 6 for a pump power of 80 mW. The peak output power obtained was 0.48 mW at 332 GHz, for 80-mW pump power, while the best efficiency occurred at this frequency with 30-mW pump power, being 0.75 percent.

VI. Conclusion

A new approach to the design of higher order varactor frequency multipliers for millimeter and submillimeter wavelengths has been described. The device uses waveguide and stripline circuit elements to efficiently couple the pump power to the varactor diode. Quasi-optical structures are used to filter, match, and terminate the output and idler frequencies. A frequency tripler for the 200–280-GHz band and a tripler and quadrupler for the 260–350-GHz band have been constructed and tested. These devices show tuning bandwidth and conversion efficiency which are as good or better than currently available waveguide multiplier mounts for the same frequency ranges [4], [5], and have the added advantage of being directly compatible with the quasi-optical LO diplexing structures now widely used in millimeter wavelength heterodyne receivers.

Acknowledgment

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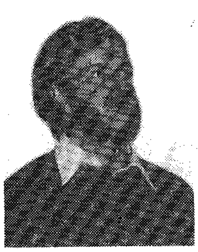

John W. Archer (M'82–SM'83) was born in Sydney, Australia in 1950. He received the B.Sc., B.E. (with first-class honors), and Ph.D. degrees from Sydney University in 1970, 1972, and 1977, respectively.

From 1974 to 1977, he was responsible for the successful development of a unique (at that time) variable baseline, phase stable, two-element interferometer for solar astronomical investigations at 100 GHz. During this period, he was associated with both Sydney University and CSIRO, Division of Radiophysics, in Australia. From 1977 to 1979, he was with the NRAO's VLA project. During this period, he was responsible for the evaluation and improvement of the performance of the overdimensioned waveguide system, as well as for the design of components for the IF section of the array. Since 1979, he has been at NRAO's Central Development Laboratory, where his main responsibility has been to coordinate the development of state-of-the-art millimeter wavelength receiver technology for use at the NRAO Kitt Peak antenna. This work has entailed the development of low-noise mixers, harmonic generators, and novel mechanical and quasi-optical structures for millimeter wave receivers.

Analysis and Design of Branch-Line Hybrids with Coupled Lines

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Abstract —**The scattering parameters of four-ports consisting of coupled lines with coupled or uncoupled connecting branches are derived in terms of the even- and odd-mode impedances and the lengths of the lines. These are used to analyze and formulate basic design procedures for the application of these structures as directional couplers including 0° and 90° 3-dB hybrids. The proposed new structures are quite compact at lower frequencies as compared to conventional uncoupled branch-line and rat-race hybrids. The results for the case of the coupled-line four-port with uncoupled branch lines also lead to closed-form expressions for the lengths and impedances of the lines required to nullify the effect of coupling between the main lines of a conventional branch-line coupler for use at higher frequencies. The measured response of the fabricated couplers is in good agreement with the theoretical predictions.**

I. Introduction

BRANCH-LINE HYBRIDS [1] are used extensively at microwave frequencies for a host of applications including phase shifters, balanced amplifiers, and in measurement systems. The major problems arising in the design of such hybrids are the junction effects, coupling between the lines, and dispersion effects. In addition, single-section hybrids are essentially narrow-band structures unless they are designed with matching sections at all the ports [2]. Methods to modify the design to incorporate the junction effects [3]–[6] and various analytical, computational, and experimental (based on heuristic logic) techniques to increase the bandwidth of such structures have also been studied [2], [7]–[9]. However, analytical methods to incorporate the effect of coupling between the lines, particularly the low-impedance lines, which can be significant at higher frequencies, are not available. In this paper, the analysis and basic design procedures for branch-line hybrids with coupled strips as shown in Fig. 1 are presented. Fig. 1(a) represents the conventional branch-line coupler at high frequencies where the physical length of the lines becomes small resulting in significant coupling between parallel lines. The proposed new four-ports of Fig. 1(b) and (c) are relatively compact structures where both the main lines and the branch lines are intentionally coupled. The scattering parameters of these four-ports consisting of coupled lines with coupled or uncoupled branches are derived in terms of even- and odd-mode impedances and the lengths of various lines. These are then used to determine the impedances and lengths of various lines such that the four-port is matched and the division of power between the coupled and the direct ports is in accordance with a desired specified value (e.g., equal for a 3-dB coupler) at the center frequency of the coupler. It is shown that the effect of coupling between the low-impedance lines which may become significant at higher frequencies (e.g., X-band) for conventional branch-line couplers [1] can be compensated by modifying the length and impedances of the lines. In addition, it is shown that compact hybrids consisting of coupled strips with folded branches having frequency response characteristics similar to the conventional branch-

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