IEICE TRANS. ELECTRON., VOL. E81-C, NO. 6 JUNE 1998

PAPER Special Issue on Microwave and Millimeter-Wave Module Technology

Design of a Quasi-Optical Oscillator Using a Grooved Mirror with a HEMT Array

Jongsuck BAE[†,††a)], Satoru SUGAWARA[†††], *Members*, Fumikazu TAKEI[†], *Nonmember*, Hiroshi KIKUYAMA[††††], *Member*, Tetsu FUJII[†,††], *Student Member*, *and* Koji MIZUNO[†,††], *Member*

SUMMARY An equivalent circuit for designing a coherent power combiner using a quasi-optical resonator has been developed. In the resonator, large numbers of devices (HEMT, HBT, etc.) are arrayed two dimensionally and mounted on a surface of a metal grooved-mirror. A newly developed equivalent circuit for the resonator has been constructed using a transmission-line model. Experiments performed at Ku-band have shown that oscillation frequencies in a 3×3 HEMT array oscillator can be predicted with errors of less than 1% by using this equivalent circuit.

key words: *quasi-optics, coherent power combining, metal grooved-mirror, Fabry-Perot resonator*

1. Introduction

Solid-state devices such as Gunn diodes and FET's are useful radiation sources in the microwave region because of their low power consumption, compact of size, and ease of handling. However, in the short-millimeter wave region, these devices can produce only low power signals, typically several mW or less. In addition, the output is noisy because of the low Q-value of conventional waveguide resonators due to large rf losses.

One solution to these problems is to employ coherent power combining for these devices by mounting them in a quasi-optical resonator (Fabry-Perot resonator). A quasi-optical resonator can achieve a high Q-value at frequencies in the submillimeter wave region [1], and its quasi-optical structure allows for a large number of devices to be mounted in the resonator.

In 1988, we proposed a Fabry-Perot resonator with a metal grooved-mirror as a quasi-optical resonator for millimeter and submillimeter wave oscillators (refer to Fig. 1) [2]. These grooved-mirror type resonators [3] have been applied to combine power for several kinds of devices such as Gunn diodes [4], HEMT's [5], and resonant tunneling diodes [6]. In this paper, we describe a newly developed equivalent circuit which can be used to design a grooved-mirror type resonator using three-terminal devices. Experimental verification of the equivalent circuit are given with a 3×3 HEMT array oscillating at Ku-band.

2. Oscillator Configuration

Figure 1 shows the configuration of a grooved-mirror type quasi-optical oscillator with transistor devices, in our case HEMT's. The oscillator consists of a metal grooved-mirror and a concave mirror. A number of HEMT's are arrayed two dimensionally and are mounted on the surface of the grooved-mirror. Radiation power from each of the HEMT's is combined coherently and spatially in the resonator. The movable back shorts are adjusted to match impedances between the HEMT's and the resonator. DC bias to the HEMT's is provided through metal plates forming the grooved-mirror. Bias ribbons to each HEMT act as an antenna array and couple the HEMT's to the resonator.

This configuration of the oscillator has several advantages such as accommodating a large number of elements, simple bias circuitry, and providing a means for adjusting the circuit impedance by the groove dimensions. Furthermore, the metal grooved-mirror makes excellent heat sink and thus ensures stable oscillation for a large number of array elements. These features imply that the grooved-mirror type oscillator is suitable as a high power source for millimeter and submillimeter wavelengths.

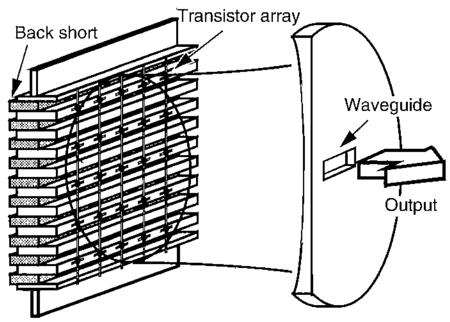


Fig. 1 A grooved-mirror type quasi-optical oscillator.

Manuscript received October 27, 1997.
Manuscript revised January 10, 1998.

† The authors are with Research Institute of Electrical Communication, Tohoku University, Sendai-shi, 980-8577 Japan.
†† The authors are also with Photodynamics Research Center, The Institute of Physical and Chemical Research (RIKEN), Aoba-ku, Sendai-shi, 980-0868 Japan.
††† The author is with General Electronics Research and Development Center, Ricoh Co., Ltd., Natori-shi, 981-1241 Japan.
†††† The author is with Advanced Technology Research Laboratories, Electronics Components & Processing Research Department, Sumitomo Metal Industries, Ltd., Amagasaki-shi, 660-0856 Japan.
a) E-mail: bae@riec.tohoku.ac.jp

3. Equivalent Circuit

An equivalent circuit for designing the grooved-mirror power combiner was developed. To make the circuit, it was assumed that the array with an infinite number of HEMT's had a symmetric structure and all of the HEMT's were identical and oscillated in phase in the resonator. It was also assumed that electromagnetic fields in the resonator were uniformly distributed and were polarized in perpendicular to the grooves. From these assumptions, a unit cell was defined as shown in Fig. 2(a), in the same way as for modeling a grid oscillator [7]. The unit cell had dimensions of $2D \times s$, where D is a period of the grooves and s is a spacing between the HEMT's. For the unit cell, an equivalent waveguide was also defined, which had boundaries consisting of electric walls on the top and bottom, and magnetic walls on the sides. An equivalent circuit for the resonator was constructed by applying a transmission-line model to the equivalent waveguide, and is shown in Fig. 2(b).

In Fig. 2(b), one section of transmission line is used to model the inside of the groove, and another section is used for the outside. Since these transmission lines are modeled as parallel-plate transmission lines in the TEM_{00} mode, the characteristic impedances are simply determined by the dimensions of their cross-sections. These are $Z_G=Z_M \times d/2D$ for the inside of the groove and $Z_M=Z_0 \times 2D/s$ for the outside, where Z_0 (~377Ω) is the characteristic impedance of free space and d is the groove height. The other parameters are introduced as lumped elements. These are fringing capacitance C_f arising from the discontinuity at the surface of the grooved-mirror, lead inductance L_l, parasitic capacitance C_l at the gate and drain of HEMT, and a center-tapped transformer connected between the waveguides on the inside and outside of the groove. The impedance Z_L is used to account for a coupling loss at the output mirror as well as other rf-losses in the resonator.

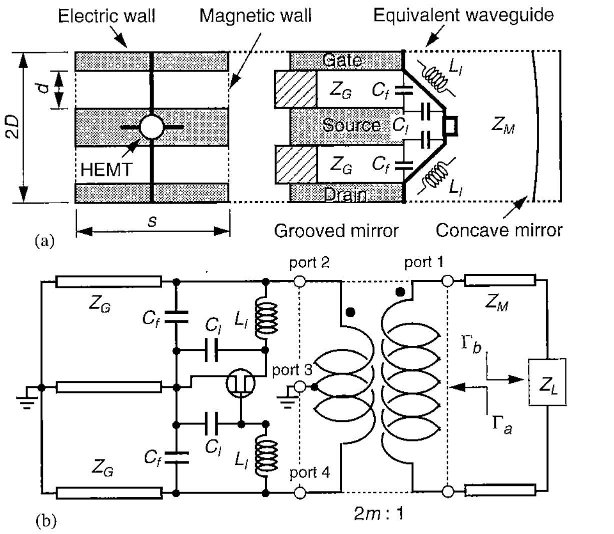


Fig. 2 (a) The unit cell and (b) the equivalent circuit for the HEMT array in the resonator.

The fringing capacitance, C_f, changes depending on the period and height of the grooves, and is determined by analyzing electromagnetic field distributions in the metal grooved-mirror [1]. C_l is estimated simply from the size of the metal lead of the gate (or the drain) on the metal source plate (Fig. 2(a)). L_l is determined for the lead length and width by using the induced EMF method [8], [9] (see Appendix). In Fig. 2(b), the transformer with four ports is introduced to distribute rf-power to the insides of the two grooves and the outside without power losses. Its S-parameters [9] are given by the following equation.

$$S = \begin{bmatrix} \frac{z-2}{z+2} & \frac{2\sqrt{2}}{z+2} & 0 & \frac{-2}{z+2} \\ \frac{2\sqrt{2}}{z+2} & \frac{4-z}{3(z+2)} & \frac{2}{3} & \frac{2(z-2)}{3(z+2)} \\ 0 & \frac{2}{3} & -\frac{1}{3} & \frac{2}{3} \\ \frac{-2}{z+2} & \frac{2(z-2)}{3(z+2)} & \frac{2}{3} & \frac{4-z}{3(z+2)} \end{bmatrix}, \text{ for } z = \frac{Z_M}{Z_G}\left(=\frac{1}{m}\right)$$

where m is the winding ratio of the center-tapped transformer, which is defined as a groove voltage equal to $d/2D$ at the surface of the grooved-mirror.

The oscillation condition for a transistor in the resonator is given by $\Gamma_a \Gamma_b=1$, where Γ_a and Γ_b are reflection coefficients indicated in Fig. 2(b). Γ_a can be calculated by using the equivalent circuit along with the S-parameters for the HEMT. Since Γ_b is determined by the resonator losses, the oscillation frequencies of HEMT's in the resonator are predicted from the oscillation condition.

Though the equivalent circuit has been derived assuming an infinite HEMT array, it could be used for a practical oscillator with limited numbers of HEMT's when the HEMT array oscillates in the TEM_{00} mode of the resonator. When a fundamental Gaussian beam in the TEM_{00} mode fully covers the HEMT array in the resonator, electromagnetic fields are uniformly distributed in the unit cell, except for those in close proximity to the HEMT's. As a result, the boundary conditions in the equivalent waveguide shown in Fig. 2(a) is satisfied for a practical HEMT array.

4. Experimental Verification

Experimental verification of the equivalent circuit was carried out by comparing measured oscillation frequencies with theoretical ones in a grooved-mirror type HEMT oscillator at frequencies around 15 GHz. Figure 3 shows the oscillator configuration for experiments. The packaged HEMT's (Mitsubishi Electric co., MGF4314D) were used in the experiments. The grooved-mirror had dimensions of 130 mm × 130 mm, a period of 8 mm, and a height of 2 mm. The nine HEMT's were configured in a 3×3 array with a spacing, s, of 20 mm on the grooved-mirror (Fig. 2(a)). The HEMT's were connected to the metal plates with copper ribbons with a width of 3 mm and a thickness of 0.1 mm. Dielectric (Teflon) plates with a thickness of 0.76 mm and a width of 6 mm were used to DC insulate the drain and gate leads from a metal plate connected to the source lead at the grooved mirror. The metal concave mirror had a diameter of 200 mm and a radius of curvature of 400 mm. The output power and frequencies were measured through a rectangular waveguide with dimensions of 23.4 mm × 9.3 mm at the center of the concave mirror, by using an HP-8563A spectrum analyzer.

Figure 4 shows the measured frequency spectra for grooved-mirror type oscillators with (a) one HEMT and (b) nine HEMT's. The oscillation frequencies for both oscillators were about 16 GHz. The output power and a C/N ratio at

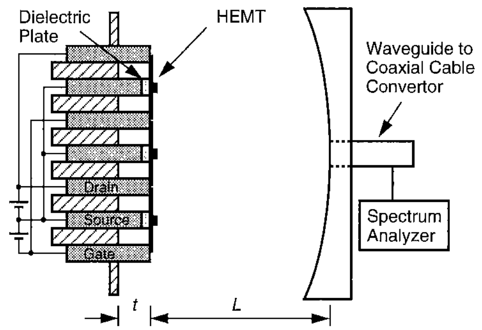


Fig. 3 Experimental setup at Ku-band.

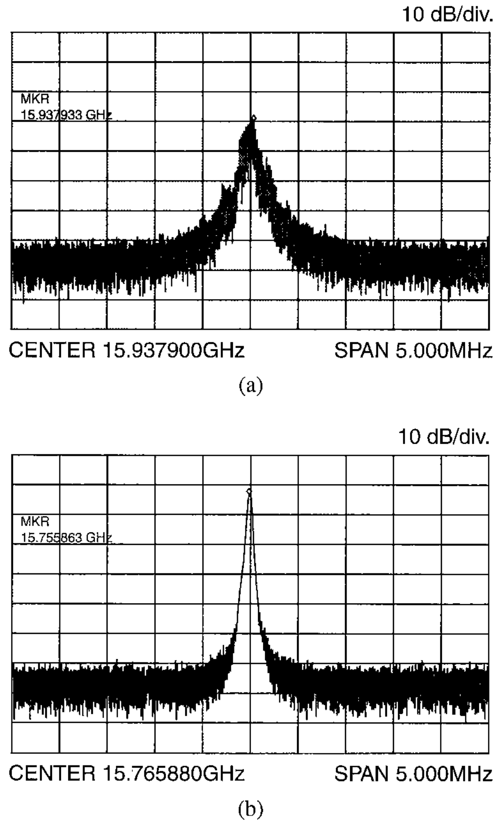


Fig. 4 Measured frequency spectra for the oscillators with (a) one HEMT and (b) nine HEMT's.

a 100 kHz offset have increased for the oscillator with nine HEMT's by 16.5 dB and 25 dB/Hz, respectively. These results show that the radiation power from each of the nine HEMT's has been coherently combined in the resonator.

Figure 5(a) compares the theoretically calculated and measured oscillation frequencies for the nine HEMT's oscillator as a function of the resonator length. The measured corresponding output power is also indicated in Fig. 5(b). The black circles indicate the measured oscillation frequencies and output power in the fundamental TEM_{00} mode, and the white circles are those in higher order modes, such as TEM_{10} mode. In the experiments, the groove depth, t, in Fig. 3 was adjusted to give the largest tuning frequency range in the TEM_{00} mode of oscillation, and was about 4.1 mm.

In Fig. 5(a), the theoretical frequencies are plotted for an equivalent circuit with a unit cell size of 16 mm × 20 mm and the small signal *S*-parameters of the HEMT's. The *S*-parameters were determined at frequencies between 2 GHz and 18 GHz using a microstrip line circuit. Γ_b of −0.76 which was determined by measuring the loaded Q-value of the resonator, was used for calculation. The calculation parameters, $C_f = 0.049$ pF, $C_l = 0.15$ pF, and $L_l = 0.49$ nH, were used and estimated from the physical dimensions of the resonator. Since the Teflon plate was very thin compared to the wavelength, its dielectric constant was only taken into account for calculating the parasitic capacitance, C_l.

From Fig. 5(a), it is seen that the theoretical results agree with the measured ones with errors of less than 1%, except for the oscillation frequencies in the higher order modes. It should be noted that no fitting was made for the calculation

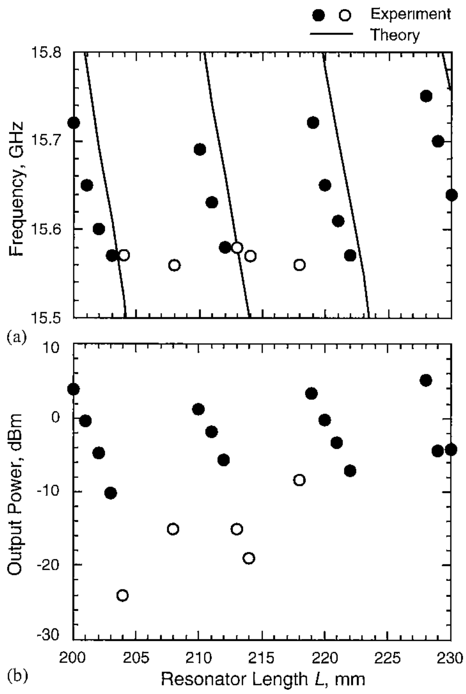


Fig. 5 (a) Measured oscillation frequencies and (b) output power for the oscillator with nine HEMT's, as a function of the resonator length. The solid lines in (a) indicate the theoretically calculated frequencies.

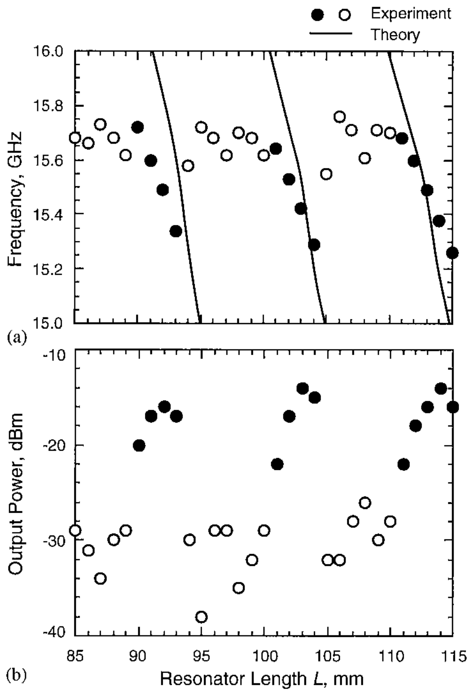


Fig. 6 (a) Measured oscillation frequencies and (b) output power for the oscillator with one HEMT, as a function of the resonator length. The solid lines in (a) indicate the theoretically calculated frequencies.

parameters. These results indicate that the equivalent circuit for the oscillator can be used for a small array. This good agreement between the theory and experiment was most likely due to the TEM_{00} mode of operation as described in Sect. 3. For the resonator with a 3×3 HEMT array used in this experiment, a beam with a diameter of about 70 mm was produced at the surface of the grooved-mirror when operating in the TEM_{00} mode. This beam covered the entire HEMT array.

In order to confirm the applicability of the equivalent circuit, we compared the calculated and measured oscillation frequencies for the oscillator with a single HEMT. These results are shown in Fig. 6. Again, the black circles indicate the oscillation frequencies and output powers in the TEM_{00} mode. In the experiments, a concave mirror with a radius of curvature of 200 mm was used to reduce diffraction loss. For this resonator operating in the TEM_{00} mode, the beam diameter at the surface of the grooved mirror was about 50 mm. In the equivalent circuit for this oscillator, a square unit cell with dimensions of 16 mm × 16 mm for the single HEMT was defined and used for calculation, which gave the characteristic impedance Z_M equal to Z_0 in free space in the equivalent circuit. The calculation parameters were $C_l = 0.22$ pF and $L_l = 0.35$ nH. The fringing capacitance, C_f, was the same as the previous one. From Fig. 6(a), it is seen that the theory can predict the oscillation frequencies within errors of about 3%. These results imply that the boundary conditions assumed for the unit cell in Fig. 2(a) hold even for a single HEMT when operating in the TEM_{00} mode.

5. Conclusion

An equivalent circuit for a grooved-mirror type resonator for three-terminal devices has been developed to theoretically design the resonator. The equivalent circuit was constructed by using a transmission-line model for a unit cell of the transistor array. The theoretical circuit for the resonator was verified experimentally at Ku-band. The experimental results show that the equivalent circuit is applicable even in the case of a resonator with only a single transistor, when using a square unit cell and when operating in the TEM_{00} mode.

Acknowledgment

The authors would like to acknowledge Dr. Y. Iyama, Mr. M. Hieda, and Dr. M. Nakayama at Mitsubishi Electric co. for providing us with the HEMT's and for assistance with their technical treatment. The research project was partially supported by a Grant-in-Aid for Scientific Research (B), 09450134, 1997, from the Ministry of Education, Science, Sports and Culture, Japan.

References

[1] K.Mizuno and S.Ono, "The ledatron," in Infrared and Millimeter Waves, vol. 1, K. Button, ed., Ch. 5, 1979.

[2] K.Mizuno, T.Ajikata, M.Hieda, and M.Nakayama, "Quasi-optical resonator for millimeter and submillimeter wave solid-state sources," Electron Lett., vol.24, no.13, pp.792-793, 1988.

[3] H.Kondo, M.Hieda, M.Nakayama, T.Tanaka, K.Osakabe, and K.Mizuno, "Millimeter and submillimeter wave quasi-optical oscillator with multi-elements," IEEE Trans. Microwave Theory & Tech., vol.40, no.5, pp.857-862, 1992.

[4] J.Bae, Y.Aburakawa, H.Kondo, T.Tanaka, and K.Mizuno, "Millimeter and submillimeter wave quasi-optical oscillator with gun diodes," IEEE Trans. Microwave Theory & Tech., vol.41, no.10, pp.1851-1855, 1993.

[5] J.Bae, M.Akaishi, Y.Aburakawa, and K.Mizuno, "Design of a millimeter wave quasi-optical oscillator," in Digest 18th Int. Conf. Infrared and Millimeter Waves, Colchester, U. K., pp.474-475, 1993.

[6] T.Fujii, H.Mazaki, F.Takei, J.Bae, M.Narihiro, T.Noda, H.Sakaki, and K.Mizuno, "Coherent power combining of millimeter wave resonant tunneling diodes in a quasi-optical resonator," 1996 IEEE MTT-S Int. Microwave Symp. Dig., pp.919-922.

[7] R.M.Weikle II, M.Kim, J.B.Hacker, M.P.DeLisio, Zoya B.Popovic, and D.B.Rutledge, "Transistor oscillator and amplifier grids," Proc. IEEE, vol.80, no.11, pp.1800-1809, 1992.

[8] R.L.Eisenhart and P.J.Khan, "Theoretical and experimental analysis of a waveguide mounting structure," IEEE Trans. Microwave Theory & Tech., vol.MTT-19, no.8, pp.706-719, 1971.

[9] R.M.Weikle II, "Quasi-optical planar grids for microwave and millimeter-wave power combining," Ph. D. Dissertation, California Institute of Technology, 1992.

Appendix

The impedance of the metal leads of HEMT at the surface of the metal grooved-mirror was estimated by using the induced Electro-Motive Force (EMF) method. In this method, the lead

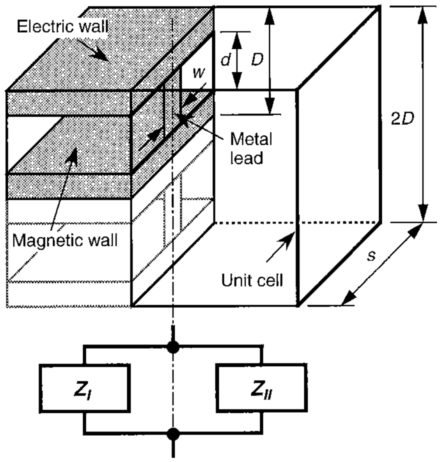


Fig. A · 1 Calculation model for the lead impedance in the unit cell.

impedance is calculated by assuming a uniform current distribution in the metal lead.

Figure A · 1 shows the calculation model for the lead impedance in an equivalent waveguide with electric and magnetic walls. The equivalent waveguide consists of a waveguide, $\mathbf{W}_\mathrm{I}$, with dimensions of $d{\times}s$ for the groove, and a waveguide, $\mathbf{W}_\mathrm{II}$, with dimensions of $2D{\times}s$ for free space. For ease of calculation, we have assumed that the same metal leads are placed in the two waveguides, $\mathbf{W}_\mathrm{I}$ and $\mathbf{W}_\mathrm{II}$, both having infinite length. A metal lead with a width of w and a thickness of zero is also assumed. The lead impedances, Z_I in $\mathbf{W}_\mathrm{I}$ and Z_{II} in $\mathbf{W}_\mathrm{II}$ are calculated separately using the EMF method, and then the lead impedance in the equivalent waveguide is determined as the parallel connection of Z_I and Z_{II} as shown in Fig. A · 1. Since the calculation procedures for impedances, Z_I and Z_{II}, are almost the same as Weikle's [9], only the calculation results are presented here.

The impedance, Z_{lead}, in the metal lead is given by the following equations,

$$Z_{lead}=Z_I//Z_{II}$$

$$Z_I=\frac{b}{a}Z_0^{TEM}+\frac{2b}{a}\sum_{m=1}^{\infty} Z_{m0}^{TE}\cos^2\left(\frac{m\pi}{2}\right)\mathrm{sinc}^2\left(\frac{m\pi w}{2a}\right)$$

for $a=s$ and $b=d$,

$$Z_{II}=\frac{d^2}{ab}Z_0^{TEM}+\frac{2d^2}{ab}\sum_{n=1}^{\infty} Z_{0n}^{TM}\cos^2\left(\frac{3n\pi}{4}\right)\mathrm{sinc}^2\left(\frac{n\pi d}{2b}\right)$$
$$+\frac{d^2}{ab}\sum_{m=1}^{\infty}\sum_{n=1}^{\infty}\frac{4k_y^2}{k_c^2}Z_{mn}^{TM}\cos^2\left(\frac{m\pi}{2}\right)\cos^2\left(\frac{3n\pi}{4}\right)\mathrm{sinc}^2\left(\frac{m\pi w}{2a}\right)$$
$$\times\mathrm{sinc}^2\left(\frac{n\pi d}{2b}\right)+\frac{2d^2}{ab}\sum_{m=1}^{\infty}Z_{m0}^{TE}\cos^2\left(\frac{m\pi}{2}\right)\mathrm{sinc}^2\left(\frac{m\pi w}{2a}\right)$$
$$+\frac{d^2}{ab}\sum_{m=1}^{\infty}\sum_{n=1}^{\infty}\frac{4k_x^2}{k_c^2}Z_{mn}^{TE}\cos^2\left(\frac{m\pi}{2}\right)\cos^2\left(\frac{3n\pi}{4}\right)$$
$$\times\mathrm{sinc}^2\left(\frac{m\pi w}{2a}\right)\mathrm{sinc}^2\left(\frac{n\pi d}{2b}\right)$$

for $a=s$ and $b=2D$.

where $\mathrm{sinc}(x)=\sin(x)/x$. The characteristic wave impedances, Z_0^{TEM}, Z_{mn}^{TM}, and Z_{mn}^{TE} for TEM_{00}, TM_{mn}, and TE_{mn} modes in the waveguides, $\mathbf{W}_\mathrm{I}$ and $\mathbf{W}_\mathrm{II}$, are given by the following equations,

$$Z_0^{TEM}=\sqrt{\frac{\varepsilon}{\mu}}\ (\cong 377\Omega),\quad Z_{mn}^{TM}=\frac{k_z}{\omega\varepsilon},\quad Z_{mn}^{TE}=\frac{\omega\mu}{k_z}$$

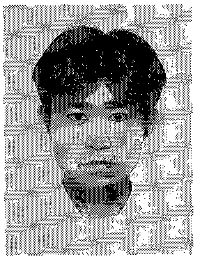
$$k_x=\frac{m\pi}{a},\quad k_y=\frac{n\pi}{b},\quad k_c^2=k_x^2+k_y^2$$

$$k_z=\sqrt{\omega^2\mu\varepsilon-k_c^2}$$

where m and n are integers, ε and μ are the permittivity and permeability of free space, respectively, and ω is the angular frequency.

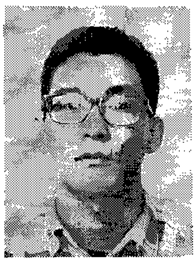
Jongsuck Bae was born in Nagoya, Japan, in 1953. He received the B. S. degree in electric engineering from Korean University, Tokyo, in 1976. He joined the Research Institute of Electrical Communication, Tohoku University, Sendai, Japan, in 1977, where he has been Associate Professor since 1992. He received the D.Eng. degrees in electronic engineering from Tohoku University in 1990. His research has been in developing oscillators, modulators, compact electron beam devices, and in applications of near field optics to quasi-optical components, in millimeter and submillimeter wave frequency regions. Dr. Bae is a member of IEEE, and the Japan Society of Applied Physics.

Satoru Sugawara was born in Miyagi, Japan, in 1965. He received the B.S. degree in physics from Tohoku University, Sendai, in 1989. He joined Ricoh research institute of general electronics, in 1989. He is currently with general electronics R&D center, Ricoh Co. Ltd.. His current work is in developing on mm-wave antennas and their applications. Mr. Sugawara is a member of IEEE.

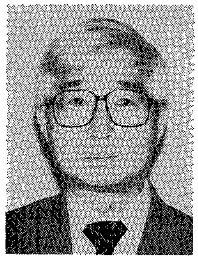


Fumikazu Takei was born in Kanagawa, Japan, in 1971. He received the B.Eng. and M.Eng. degrees in electronic engineering from Tohoku University, Sendai, in 1995 and 1997, respectively. He joined NEC co., in 1997. He is currently with Test Systems Development Department, NEC co.. His current work is in developing on Wafer Probing Machine.

Hiroshi Kikuyama was graduated from The National Defense Academy (NDA) in 1989. His undergraduate studies were focused on Piezoelectric Ceramics. In 1989, he joined Kyocera Corporation and was engaged in development of Electronic Ceramics for Multi-Layer Ceramic Capacitor (MLC) until 1991. In 1991, he joined Sumitomo Metal Industries (SMI) and was engaged in development of RF components (filters, resonator, capacitor, etc.) and EMC Components (noise filter) for 5 years. From 1996 to 1997 he was studying about Quasi-Optical Power Combining Techniques (Quasi-Optical Oscillator with HEMT's) at Research Institute of Electrical Communication in Tohoku University. Since 1997, he has been engaged in research and development of Millimeter-Wave Components (antenna, MMIC's package, filter, low-noise amplifier) at Advanced Technology Research Laboratories in SMI.



Tetsu Fujii was born in Tokyo, Japan in 1969. He received B.E. and M.E. degrees in electronic engineering from Tohoku University, Sendai, Japan in 1993 and 1995, respectively. He is a student of the doctor course of electronic engineering in Tohoku University and his research interest is in coherent power combining of solid state devices using a quasi-optical resonator in millimeter and submillimeter wave regions.



Koji Mizuno was born in Sapporo, Japan on July 17, 1940. He received the B.Eng., M.Eng., and D.Eng. degrees in electronic engineering from Tohoku University, Sendai, in 1963, 1965, and 1968, respectively. In 1968, He joined the Department of Electronic Engineering, Faculty of Engineering, at Tohoku University. He was appointed Associate Professor at Research Institute of Electrical Communication in 1972, and since 1984 has been a Professor of Electron Devices there as well as in Department of Electronic Engineering. In 1973 he spent a one-year sabbatical leave at Queen Mary College, University of London, under the sponsorship of SRC (Science Research Council, United Kingdom), and in 1990 spent a six-months sabbatical leave at both the California Institute of Technology, Pasadena, and Queen Mary College, London, under the sponsorship of Monbusho (Ministry of Education, Science, Sports and Culture, Japan). He is interested in the millimeter and submillimeter wave region of the electromagnetic spectrum, and his current work is in detection and generation technologies, and their applications within the region. Dr. Mizuno is member of the Institute of Electronics, Information and Communication Engineers, the Institute of Electrical Engineers of Japan, and the Japan Society of Infrared Science and Technology. In 1984 he received the 17th Kagaku Keisoku Shinkokai (Scientific Measurement) award. In 1990 he became a team leader at the Photodynamics Research Center (the Institute of Physical and Chemical Research), Sendai, and is now running a laboratory for submillimeter wave research there, as well as one at Tohoku University.