A Periodic Branching Filter for Millimeter-Wave Integrated Circuits

TAKAO ITANAMI

Abstract—**A number of passive and active devices using dielectric waveguides have been developed and find various applications in integrated circuits at the millimeter optical-frequency range.**

The design, theoretical considerations and experimental findings of a periodic branching filter using rectangular dielectric waveguides are described in this paper.

Low insertion loss for the periodic branching filter with 850-MHz 3-dB bandwidth, less than 1.0 dB, is achieved in the frequency range from 77 to 85 GHz. Measured results are in good agreement with theoretical calculations.

I. Introduction

MILLIMETER-WAVE integrated circuits, which use a dielectric waveguide have been emerging for the past several years. Both dielectric rectangular waveguides and insular dielectric waveguides seem to be suitable for millimeter-wave integrated circuits. One of the most important characteristics required for the waveguide composing the circuit component is low loss. However, insular dielectric waveguides have essentially higher loss than dielectric rectangular waveguides due to the presence of the metal ground plane. It thus appears that dielectric rectangular waveguides are more suitable for millimeter-wave integrated circuits.

Circuit components using dielectric rectangular waveguides, however, have not been sufficiently investigated. A channel diplexer, one of the circuit components using a dielectric rectangular waveguide, is described in [1]. This ring-type channel diplexer is composed of ring resonators and input–output guides. The insertion loss of this filter is greatly affected by the resonator loss, as the resonance point of the resonator is placed at the center of the passband.

To obtain a ring-type diplexer with a wide 3-dB bandwidth, the resonator length must be shortened. That is, the bending radius of ring resonator must become small. This causes an increase of insertion loss due to the increase in ring resonator radiation loss, which is very difficult to avoid. For this reason, it is difficult to obtain a low-loss ring-type diplexer using a dielectric waveguide with a relatively wide 3-dB bandwidth of about 1 percent.

A periodic branching filter, whose amplitude transmission characteristics versus frequency vary periodically is well known and has been described in several papers [2]–[5]. A periodic branching filter with a traveling-wave resonator has been applied in band-splitting filters for satellite use [6].

When using this filter as a channel diplexer that has a relative 3-dB bandwidth of about 1 percent, a long-length traveling-wave resonator is needed. As a result, the bend radius of the resonator for this filter becomes large. This causes a decrease in the radiation loss form the resonator when the filter is made of dielectric waveguide.

Another feature of the periodic filter is that they essentially have low-loss characteristics as the resonant frequency of the resonator is placed at the edge of each passband, and the passband loss for this filter is not degraded by the resonance. Therefore, this filter is considered to be suitable for low-loss constructions using dielectric waveguides.

This paper describes the design method and theoretical considerations of the periodic filter that uses a rectangular dielectric waveguide. Experimental results are also described for the periodic filter designed as a 80-GHz band channel branching filter.

II. Structure and Design

A periodic branching filter with a traveling-wave resonator is composed of three directional couplers, connecting waveguides for phase adjustment, and a traveling wave resonator. Two of the three couplers are 3-dB hybrids. The other one is used to couple the resonator to one of the connecting waveguides. An equivalent circuit for this filter is shown in Fig. 1.

Theoretical transmission characteristics for this filter are given by the equations in the Appendix. Phase and amplitude characteristics for this filter are shown in Fig. 2. It is assumed in the calculations that the three directional couplers are ideal and that the waveguides are lossless. In Fig. 2, the axis of the abscissa is the electrical length of the resonator and the axis of the ordinate is the phase or amplitude. The straight solid line in Fig. 2(a) indicates the phase difference between two connecting waveguides, while the winding solid line and broken line indicate the phase variations of the resonator, including the coupler, at $k_0 = 2\sqrt{2}/3$ and at $k_0 = 1$, respectively. When k_0 equals to

Manuscript received December 8, 1980; revised April 16, 1981.

The author is with the Yokosuka Electrical Communication Laboratory, Nippon Telegraph and Telephone Public Corporation, Yokosuka-shi, Kanagawa-ken, 238-03 Japan.

0018-9480/81/0900-0971$00.75 ©1981 IEEE

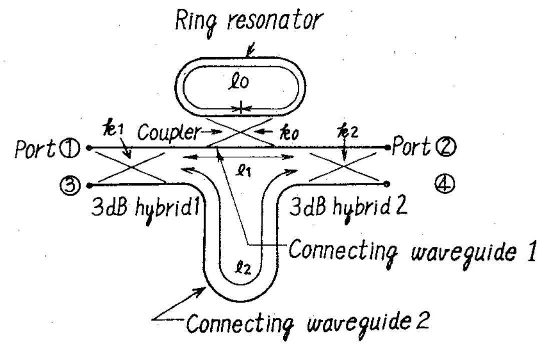

Fig. 1. Equivalent circuits for periodic branching filter

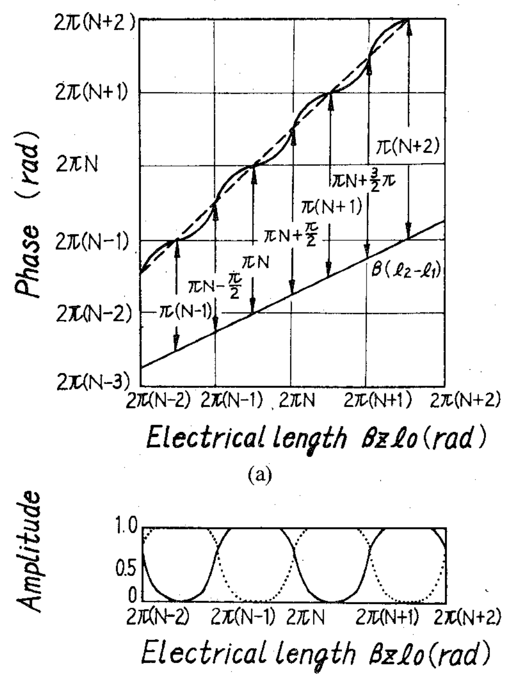

Fig. 2. Phase and amplitude characteristics for periodic branching filter. (a) Phase. (b) Amplitude.

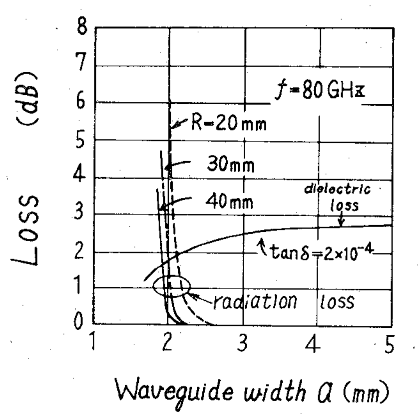

Fig. 3. Loss characteristics of rectangular dielectric waveguide.

$2\sqrt{2}/3$, inclinations of the resonator phase ϕ and $\beta_z(l_2-l_1)$ coincide at each antiresonance point for the resonator (i.e., $\beta_z l_0=(2n+1)\pi$). The maximally flat amplitude response can then be obtained. The amplitude characteristics of this filter is shown in Fig. 2(b). This figure also shows that the resonance point of the resonator is placed at the edge of each passband. For that reason, this filter's insertion loss is essentially low in contrast to the usual diplexer using resonators, in which the resonance point is placed at the center of the passband.

In designing this filter, the waveguide material and dimensions must first be determined. Poly-tetra-fluoroethylene (PTFE) was chosen for the waveguide material of this filter, because of its low-loss property and ease of fabrication. Assuming that the dielectric constant and loss tangent of PTFE are 2.0 and 2×10^{-4}, respectively, theoretical radiation loss and dielectric loss can be calculated [7], [8].

The calculated results of radiation losses from a 2π radian bend and dielectric loss per meter are plotted in Fig. 3. They are shown against the waveguide width a for a square dielectric waveguide. The dielectric waveguide mode used for the calculation is the dominant E_{11}^{y} mode. Theoretical calculation indicates that radiation loss increase rapidly with decreasing waveguide width from 3 to 2 mm, when the bend radius R varies between 20 and 40 mm. The calculated results also indicate that radiation losses are negligibly small if the waveguide width is greater than 3 mm, and the bend radius R is greater than 20 mm. The dielectric loss decreases rapidly when the waveguide width decreases below 3 mm and it remains almost constant with increasing waveguide width from 3 mm.

As a consequence of these considerations, it becomes clear that the waveguide width has to be greater than 3 mm. The measured radiation loss is slightly greater than the theoretically calculated loss [7]. Therefore, the waveguide is chosen to be 3.5 mm, in order to be able to neglect radiation losses.

The propagation constant β_z along the dielectric guide is given by [8]

$$\beta_z=\left(\beta^2\epsilon_{re}-\beta_x^2\right)^{1/2} \tag{1}$$

$$\epsilon_{re}=\epsilon_r-\left(\beta_y/\beta\right)^2 \tag{2}$$

$$\beta=2\pi/\lambda \tag{3}$$

where β_x, β_y are the transverse propagation constants, ϵ_r is the relative dielectric constant of the waveguide material, and λ is the free-space wavelength.

β_x and β_y are solutions of the following transcendental equations:

$$\beta_x a=\pi-2\tan^{-1}(\beta_x/\beta_{x0}) \tag{4}$$

$$\beta_y a=\pi-2\tan^{-1}(\beta_y/\beta_{y0}) \tag{5}$$

where

$$\beta_{x0}=\left[(\epsilon_{re}-1)^2\beta_0^2-\beta_x^2\right] \tag{6}$$

and

$$\beta_{y0}=\left[(\epsilon_r-1)^2\beta_0^2-\beta_y^2\right]. \tag{7}$$

After the waveguide dimensions and material are specified, the propagation constant β_z is calculated from the equations (1) to (7) at every frequency.

If the 3-dB bandwidth B_c and the channel center frequency f_0 are given, the resonant index N of the resonator (i.e., the number of wave variations along the length of resonator) is determined by

$$N=\frac{\beta_{zN}}{\beta_{z(N+1)}-\beta_{zN}} \tag{8}$$

where β_{zN} is the propagation constant of the resonator waveguide at the frequency $(f_0-B_c/2)$ and $\beta_{z(N+1)}$ at the frequency $(f_0+B_c/2)$.

The resonator length l_0, is given by

$$l_0=N\frac{2\pi}{\beta_{zN}}. \tag{9}$$

The differential length Δl between two connecting waveguides for phase adjustment is given by

$$\Delta l=l_1-l_2=\frac{l_0}{2}-\frac{\pi}{2\beta_{zN}}. \tag{10}$$

Using these design equations, a periodic branching filter was designed under the conditions $(f_0-B_c/2)=80$ GHz and $B_c=850$ MHz. The dimensions of the designed filter are $N=80$, $l_0=234.22$ mm, $l_1=123.62$ mm, and $l_2=240$ mm.

III. Theoretical Characteristics

Transmission characteristics for the periodic branching filter described in Section II are compared with those for a conventional maximally flat type filter, in order to evaluate its characteristics. A comparison of the amplitude characteristics is presented in Fig. 4. This filter's frequency response is found to be nearly equal to that of a sixth- or seventh-order maximally flat type filter. Group delay characteristics are compared in Fig. 5. The difference between maximum and minimum delay times within this filter's 3-dB bandwidth corresponds to that for third- or fourth-order maximally flat type filters.

If the coupling factor for two 3-dB hybrids deviates from 3 dB, the transmission characteristics of the periodic filter are quite different from those for a periodic filter which has ideal 3-dB hybrids. The amplitude transmission characteristics of a periodic filter with a nonideal 3-dB hybrid is shown in Fig. 6. Here, the amplitude deviation is ±0.5 dB and $k_0=2\sqrt{2}/3$. It can be seen in Fig. 6 that no ripple arises in the transmission characteristics. Only leakage to the other port is brought about through the coupling factor deviation of ±0.5 dB. Due to this leakage to the other port, the transmission loss increases. The amount of increase, however, is small, and is less than 0.1 dB.

Relations between the amplitude coupling coefficient k_0 and the maximum leakage are shown in Fig. 7, with coupling factor deviations Δk_1, Δk_2 of the hybrids as parameters. As this figure shows the leakage increases to

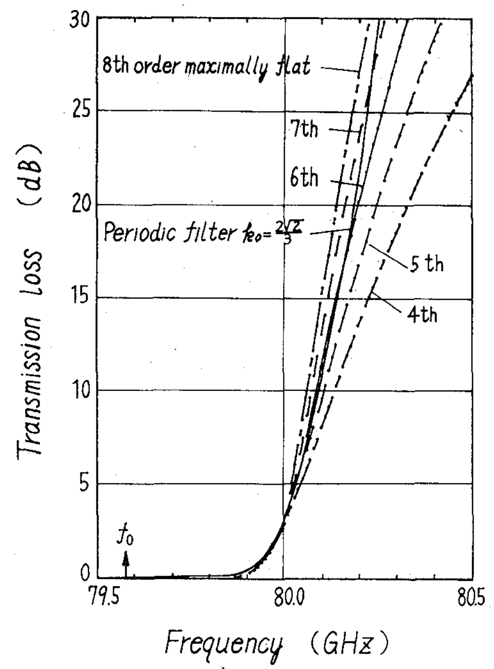

Fig. 4. Amplitude transmission characteristics comparison between designed periodic branching filter and maximally flat type filter.

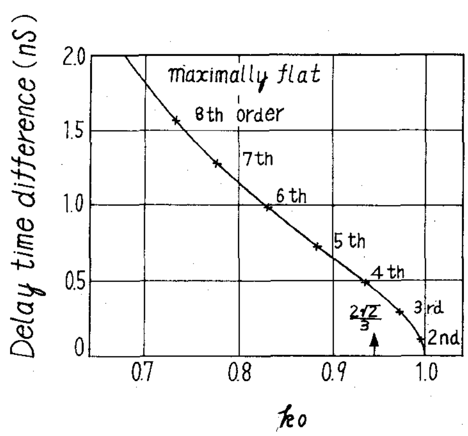

Fig. 5. Delay characteristics comparison between designed periodic filter and maximally flat type filter

about 17 dB, due to the coupling factor deviation of ± 0.5 dB under the condition that k_0 is around 0.9.

The delay time difference characteristics for this filter are shown in Fig. 8; the shading indicates the variation ranges for the delay time differences between maximum values and minimum values within the 3-dB bandwidth. The coupling factor deviation of the hybrid is within ± 0.25 and ± 0.5 dB in this figure. The delay time difference is very small when k_0 is greater than 0.9.

Theoretical amplitude transmission characteristics of the designed filter are shown in Fig. 9. The couplers are assumed to be ideal and k_0 equals to $2\sqrt{2}/3$.

The channel center frequency deviation from the nominal value ($80.425 \pm 0.825n$ GHz ($n=1,2,\cdots$)) is shown in Fig. 10. The deviation is very small and is less than 4 MHz

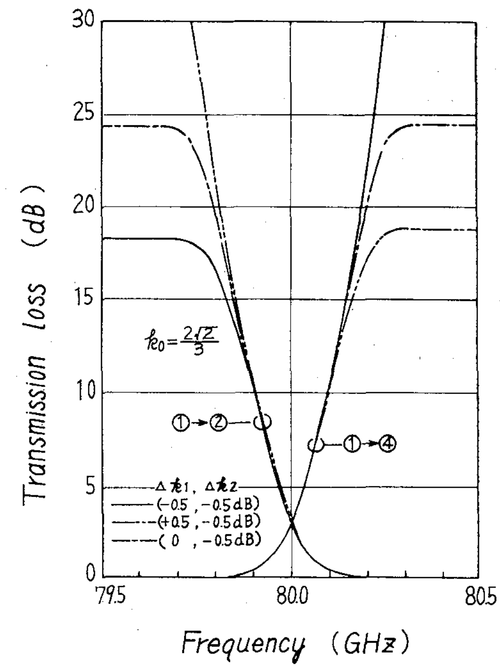

Fig. 6. Change of amplitude transmission characteristics for designed periodic filter, caused by hybrid's coupling factor deviations.

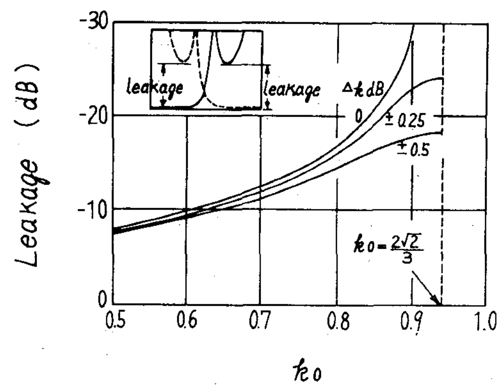

Fig. 7. Leakage increase for designed periodic filter caused by hybrid's coupling factor deviations.

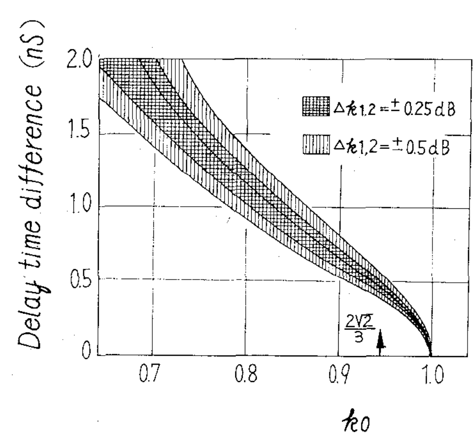

Fig. 8. Delay time difference changes for designed periodic filter, caused by hybrids' coupling-factor deviations.

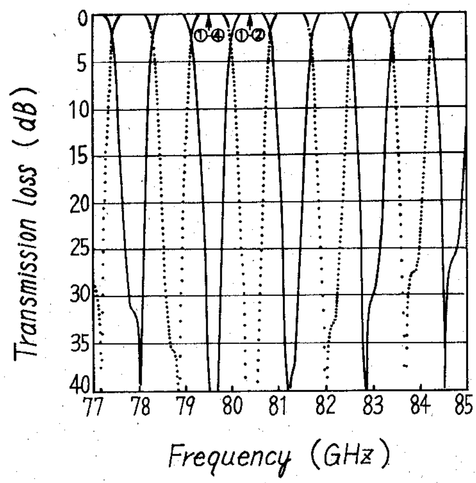

Fig. 9. Calculated transmission characteristics for designed periodic branching filter.

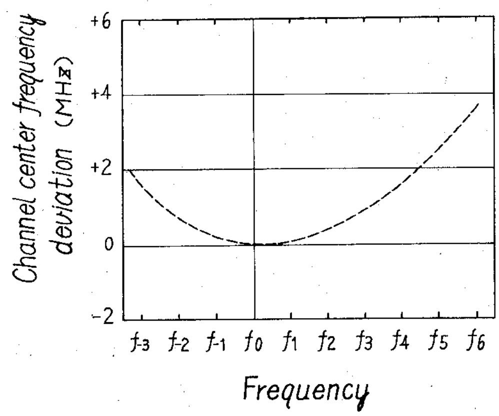

Fig. 10. Channel center frequency deviations for designed periodic-branching filter.

within the relative frequency range of 10 percent. This is due to the low dispersion of the dielectric waveguide.

IV. Characteristics of the Fabricated Periodic Filter

The transmission loss for the E_{11}^{y} mode in a dielectric waveguide with a cross section of 3.5 mm×3.5 mm was measured by a transmission method. A conventional rectangular horn was employed to launch the E_{11}^{y} mode onto the dielectric waveguide.

Losses for two different-length dielectric waveguides were measured, and the loss of the shorter one was subtracted from that of the longer one. Thus the loss of the two horn launchers was eliminated and only the loss of the dielectric waveguide was obtained. The accuracy of the measurement was within 0.1 dB. The transmission loss was 2.5 dB/m at 80 GHz.

The fabricated periodic filter is shown in Fig. 11. This filter is composed of the same guide which was used to measure the transmission loss. The bending radius of the curved waveguide portion of this filter is chosen to be 30 mm, to limit the radiation loss to a negligible level.

Coupling characteristics of the 3-dB hybrid and the coupler, which couples the resonator to one of the connecting waveguides, are shown in Figs. 12 and 13, respectively. The coupling factor deviation of the hybrid is seen in Fig. 12 to be within ±0.5 dB at a frequency range from 77 to 85 GHz. The coupling coefficient k_0 of the coupler is seen in Fig. 13 to be between 0.942 and 0.909 at the same frequency range.

Amplitude transmission characteristics of the fabricated filter are shown in Fig. 14. The solid line and broken line indicate the transmission characteristics between ports 1 and 2, and between ports 1 and 4, respectively. The residual transmission loss (i.e., the transmission loss between ports 1 and 3) is also greater than 40 dB at the frequency range from 77 to 85 GHz. The measured transmission loss at each center frequency is between 0.85 and 1.0 dB. The calculated loss (i.e., the losses of two hybrids plus the loss calculated from equations (A1)–(A11) using the measured waveguide loss) is 0.9 dB at 80 GHz. Thus, this filter's measured transmission loss is considered to be quite acceptable.

The measured channel center frequencies are about 500

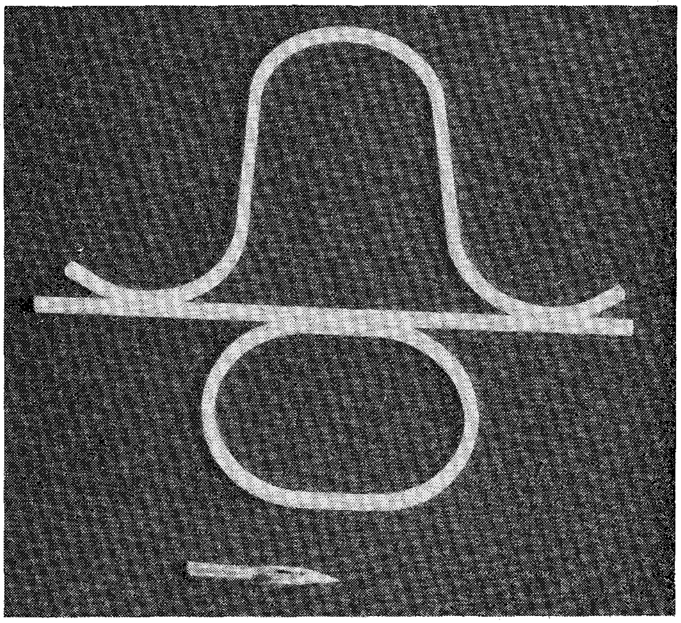

Fig. 11. Fabricated periodic branching filter.

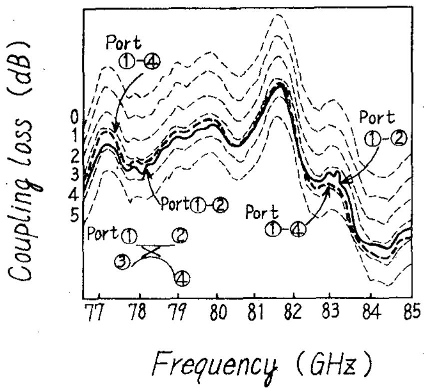

Fig. 12. Coupling characteristics of 3-dB hybrids used for fabricated periodic-branching filter.

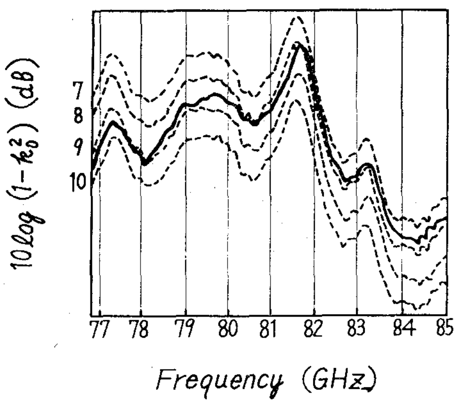

Fig. 13. Coupling characteristics between connecting waveguide and resonator of fabricated periodic-branching filter.

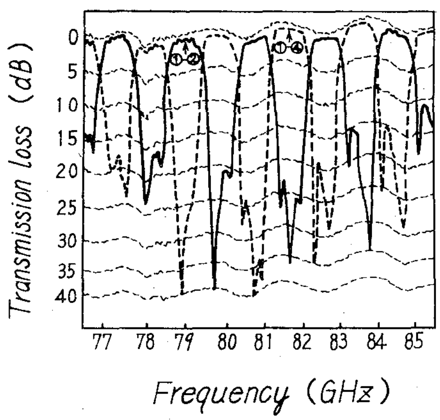

Fig. 14. Measured transmission characteristics of fabricated periodic-branching filter.

MHz higher than the design values. The measured 3-dB bandwidths are about 850 MHz and agree well with theory, as do the measured channel center frequency spacings.

The difference between the measured and theoretical channel center frequencies is caused by resonator fabrication error and the dielectric constant variation of the PTFE from the nominal value used for design. The leakage of the fabricated filter to the other output port is rather greater than predicted in Fig. 9. The reason for this increased leakage is considered to be the coupling factor deviation of the three couplers. This is indicated in Fig. 7.

V. Conclusion

The design method, theoretical characteristics, and experimental results for a periodic branching filter, using a dielectric waveguide, are presented.

The effect of the coupling-factor deviation is discussed theoretically in some detail. It is found that the transmission loss and delay characteristics are not seriously affected when the coupling factor deviation is within ± 0.5 dB and k_0 is around $2\sqrt{2}/3$. Leakage to the other output port increases seriously under the same conditions, however. Hence, when designing a periodic filter which needs low leakage levels, attention must be paid to the coupling-factor deviation of the hybrids.

A periodic filter with 850-MHz 3-dB bandwidth was designed and fabricated. A low transmission loss for this filter (less than 1.0 dB) is obtained at a frequency range from 77 to 88 GHz. The measured characteristics of this filter agree fairly well with theory and verify its essentially good transmission characteristics.

This filter can be applied to millimeter and optical integrated circuits, if waveguide material characteristics include low loss and ease of fabrication at the millimeter optical-frequency range.

Appendix

The frequency response of the periodic filter shown in Fig. 9 is given as follows:

$$A_{12}\exp(-j\Phi_{12}) \quad \text{(between port 1 and port 2)} \tag{A1}$$

$$A_{14}\exp(-j\Phi_{14}) \quad \text{(between port 1 and port 4)} \tag{A2}$$

$$A_{12}=\big[h_1^2h_2^2\exp(-2\alpha l_1)A_0^2+k_1^2k_2^2\exp(-2\alpha l_2) -2k_1k_2h_1h_2A_0\exp[-\alpha(l_1+l_2)] \cdot\cos(\phi+\beta_z(l_1-l_2))\big]^{1/2} \tag{A3}$$

$$\Phi_{12}=\tan^{-1}\left[\frac{h_1h_2\exp(-\alpha l_1)A_0\sin(\phi+\beta_z l_1)-k_1k_2\exp(-\alpha l_2)\sin\beta_z l_2}{h_1h_2\exp(-\alpha l_1)A_0\cos(\phi+\beta_z l_1)-k_1k_2\exp(-\alpha l_2)\cos\beta_z l_2}\right] \tag{A4}$$

$$A_{14}=\big[h_1^2k_2^2\exp(-2\alpha l_1)A_0+k_1^2h_2^2\exp(-2\alpha l_2)+2k_1k_2h_1h_2A_0\exp[-\alpha(l_1+l_2)]\cos(\phi+\beta_z(l_1-l_2))\big]^{1/2} \tag{A5}$$

$$\Phi_{14}=\tan^{-1}\left[\frac{-h_1k_2\exp(-\alpha l_1)A_0\cos(\phi+\beta_z l_1)-k_1h_2\exp(-\alpha l_2)\cos\beta_z l_2}{h_1k_2\exp(-\alpha l_1)A_0\sin(\phi+\beta_z l_1)+k_1h_2\exp(-\alpha l_2)\sin\beta_z l_2}\right] \tag{A6}$$

$$A_0=\left[\frac{h_0^2+\exp(-2\alpha l_0)-2h_0\exp(-\alpha l_0)\cos\beta_z l_0}{1+h_0^2\exp(-2\alpha l_0)-2h_0\exp(-\alpha l_0)\cos\beta_z l_0}\right]^{1/2} \tag{A7}$$

$$\phi=\tan^{-1}\left[\frac{(h_0^2-1)\exp(-\alpha l_0)\sin\beta_z l_0}{(1+\exp(-2\alpha l_0))h_0-(1+h_0^2)\exp(-\alpha l_0)\cos\beta_z l_0}\right] \tag{A8}$$

$$h_0=(1-k_0^2)^{1/2} \tag{A9}$$

$$h_1=(1-k_1^2)^{1/2} \tag{A10}$$

and

$$h_2=(1-k_2^2)^{1/2} \tag{A11}$$

where k_0 is the coupling coefficient between the resonator and one of the connecting guides, k_1 is the coupling coefficient of hybrid 1, k_2 is the coupling coefficient of hybrid 2, ϕ is the phase lag of the traveling-wave resonator, β_z is the phase constant of the dielectric waveguide, α is the attenuation constant of the dielectric waveguide, l_0 is the length of the resonator, and l_1, l_2 are the lengths of the two connecting waveguides.

The group delay characteristics of this filter are given by

$$\frac{d\Phi_{12}}{d\omega}=\frac{X_2}{Y_2} \quad \text{(between port 1 and port 2)} \tag{A12}$$

$$\frac{d\Phi_{14}}{d\omega}=\frac{X_4}{Y_4} \quad \text{(between port 1 and port 4)} \tag{A13}$$

$$\begin{aligned}X_2=&h_2^2h_2^2\exp(-2\alpha l_1)A_0^2\left(\frac{d\phi}{d\omega}+l_1\frac{d\beta_z}{d\omega}\right)\\&+k_1^2k_2^2\exp(-2\alpha l_2)l_2\frac{d\beta_z}{d\omega}\\&+h_1^2h_2^2k_1^2k_2^2\exp[-\alpha(l_1+l_2)]A_0\\&\cdot\cos[\phi+\beta_z(l_1-l_2)]\left[\frac{d\phi}{d\omega}+(l_1+l_2)\frac{d\beta_z}{d\omega}\right]\end{aligned} \tag{A14}$$

$$\begin{aligned}Y_2=&h_1^2h_2^2\exp(-2\alpha l_1)A_0+k_1^1k_2^2\exp(-2\alpha l_2)\\&+2k_1k_2h_1h_2\exp[-\alpha(l_1+l_2)]A_0\\&\cdot\cos[\phi+\beta_z(l_1-l_2)]\end{aligned} \tag{A15}$$

$$X_4 = h_1^2 k_2^2 \exp(-\alpha l_1) A_0^2 \left(\frac{d\phi}{d\omega} + l_1 \frac{d\beta_z}{d\omega} \right) + k_1^2 h_2^2 \exp(-2\alpha l_2) l_2 \frac{d\beta_z}{d\omega} + h_1 h_2 k_1 k_2 \exp[-\alpha(l_1 + l_2)] A_0 \cdot \cos[\phi + \beta_z(l_1 - l_2)] \left[\frac{d\phi}{d\omega} + (l_1 + l_2) \frac{d\beta_z}{d\omega} \right] \tag{A16}$$

$$Y_4 = h_1^2 k_2^2 \exp(-2\alpha l_1) A_0^2 + k_1^2 h_2^2 \exp(-2\alpha l_2) + 2k_1 k_2 h_1 h_2 \exp[-\alpha(l_1 + l_2)] A_0 \cdot \cos[\phi + \beta_z(l_1 - l_2)] \tag{A17}$$

$$\frac{d\phi}{d\omega} = \frac{X}{Y} l_0 \frac{d\beta_z}{d\omega} \tag{A18}$$

$$X = (h_0^2 - 1)\left[h_0 \exp(-\alpha l_0)(1 + \exp(-2\alpha l_0)) \cdot \cos\beta_z l_0 - (h_0^2 + 1)\exp(-2\alpha l_0) \right] \tag{A19}$$

$$Y = \left[(1 + \exp(-2\alpha l_0)) h_0 - (1 + h_0^2) \exp(-\alpha l_0) \cos\beta_z l_0 \right]^2 + (h_0^2 - 1)^2 \exp(-2\alpha l_0) \sin^2 \beta_z l_0 \tag{A20}$$

$$\frac{d\beta_z}{d\omega} = \frac{\lambda}{\lambda_g} \frac{1}{C} \tag{A21}$$

where λg is the guided wavelength in a dielectric waveguide, λ is the free-space wavelength, and c is the light velocity.

Acknowledgment

The author wishes to thank T. Kamata and S. Shindo of the Yokosuka Electrical Communication Laboratory, Nippon Telegraph and Telephone Public Corporation, for their valuable discussion and encouragement.

References

[1] T. Itanami and S. Shindo, "Channel dropping filter for millimeter integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 759–764, Oct. 1978.

[2] E. Kuhn and R. Pregla, "Commutating microwave channel separating filters with improved transfer characteristics," *AEU*, vol. 25, pp. 54–55, Jan. 1971.

[3] U. Unrau, "Periodic band-multiplexer for a TE_{01}^0-telecommunication system," *AEU*, vol. 25, pp. 56–57, Jan. 1971.

[4] J. Bodonyi, "Channeling filter for trunk waveguide communication at millimetric wavelength," *Marconi Rev.*, third quarter, pp. 160–192, 1973.

[5] I. A. Ravenscroft and G. P. Kidd, "Improved commutating channel separator for millimetric waveguide systems," *Electron. Lett.*, vol. 7, pp. 446–468, Aug. 12, 1971.

[6] H. Kumazawa and I. Ohtomo, "30-GHz-band periodic branching filter using a traveling-wave resonator for satellite applications," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, Aug. 1977.

[7] R. M. Knox, P. P. Toulios, and J. Q. Hawell, "Radiation losses in curved dielectric image waveguides of rectangular cross section," presented at IEEE MTT-S Int. Microwave Symp.

[8] R. M. Knox and P. P. Toulios, "Integrated circuits for the millimeter -optical-frequency range," in *Proc. Symp. Submillimeter Waves* (New York) Mar. 1970.

Takao Itanami was born in Fukuoka, Japan, on March 9, 1949. He received the B.S.I.E. and M.S.I.E. degrees, both from Kobe University, Kobe, Japan, in 1971 and 1973, respectively.

In 1973 he joined the Electrical Communication Laboratory, Nippon Telegraph and Telephone Public Corporation, Yohosuka, Japan, and has been engaged in research work on waveguide circuit components for millimeter-wave transmission system and satellite communication system. Recently, his major efforts have been directed toward the applications of the dielectric waveguide.

Mr. Itanami is a member of the Institute of Electronics and Communication Engineers of Japan and the Society of Instrument and Control Engineers.