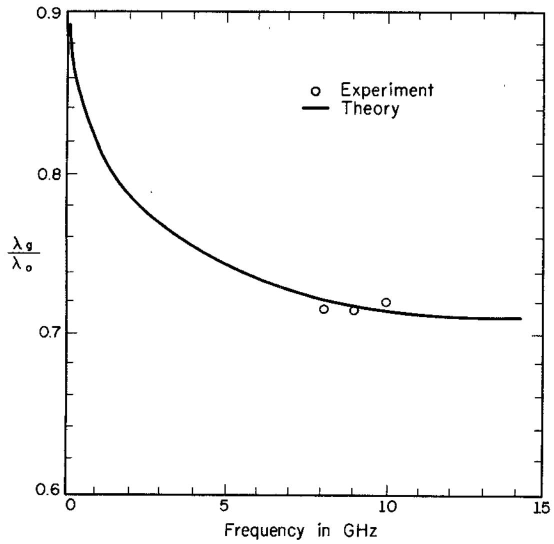

Fig. 5. Theoretical and experimental results of normalized guide wavelength of the double-layered slot line versus frequency. $w = 1.0$ mm, $d = 3.2$ mm, $h = 5.0$ mm, $\varepsilon_{r2} = 2.55$, and $\varepsilon_{r3} = 2.62$.

numerical techniques. The numerically obtained propagation constants for the dominant mode are found to be in good agreement with those obtained experimentally at X band. The computation time for evaluating a propagation constant for a given frequency is approximately 3.5 min on the G-20 computer, which is about 10 times slower than the IBM-360/75. Before this new waveguide is employed in the design of millimeter-wave IC's, extensive studies on the loss characteristics are needed.

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80-GHz-Band Low-Loss Ring-Type Channel Diplexer Using a Semicircular Electric Mode

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***Abstract*—A low-loss ring-type channel diplexer consisting of two deformed ring cavities, three directional couplers, a TE_{01} mode semicircular waveguide for the through channels, and a TE_{10} mode rectangular waveguide for the dropped channel, has been developed as a channel-dropping filter for a millimeter-wave channel multiplexing network. The structure and experimental results of the diplexer are described, the design method is also discussed. Measurements show that insertion losses of the through and dropped (coupled) channels, and VSWR are less than 0.15 dB, 0.68 dB, and 1.12, respectively, for a diplexer centered at 81.91 GHz with 3-dB bandwidth of 800 MHz. Specifically, the through channel loss is reduced by half, as compared with a conventional rectangular waveguide diplexer, owing to the low insertion loss characteristics of the TE_{01} mode semicircular waveguide. As a result, an overall loss of a channel multiplexer, in which several channel diplexers are connected in tandem, is remarkably decreased, particularly at higher frequencies covering 80–120 GHz.**

Manuscript received January 29, 1976; revised May 4, 1976.
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I. Introduction

In THE guided millimeter-wave transmission system under development in Japan, frequency ranges of 43–87 GHz [1] or more [2] are chosen in order to maintain the low loss of a millimeter-wave channel multiplexing network as well as of a circular waveguide medium. To realize such a network, several factors impose limitations on the choice of the filter structure and network construction.

There are physical limitations to the filter design at millimeter-wave frequencies because of the extremely small size. The feasibility of the filter fabrication must be considered. In addition, severe limitations exist in keeping the insertion loss low as the frequency increases. This means that hardware realization of the millimeter-wave channel multiplexing network is extremely difficult. Thus even waveguide filters of conventional structure can no longer meet the loss requirements of the guided millimeter-wave transmission system. Indeed, one of the controlling design objectives for a millimeter-wave channel diplexer is that the intrinsic losses must be kept low for both the through channel and the coupled channel. In particular, the minimization of the through-channel loss is important since many channel diplexers are connected in tandem in the repeater station. To fulfill this requirement, one might consider the utilization of low-loss modes and oversized waveguides [3]–[5].

The measured loss of a silver-plated rectangular waveguide at 80 GHz is nearly 4 dB/m (1.82 dB/m theoretically, when WR-740 is used), but that of semicircular waveguide is expected to be 0.75 dB/m (radius 3.65 mm) theoretically and 1.2 dB/m experimentally. Therefore, the oversized semicircular waveguide propagating the TE_{01} mode provides roughly a three-to-one reduction in loss, in comparison with a dominant mode rectangular guide. However, there will be some design restriction for the diplexer structure, such as location of coupling aperture, for no spurious mode should be excited in main waveguide within the operating band.

An early paper reported the feasibility of using the low-loss modes in the conventional ring-type channel diplexer [5]. In this paper, the design of a channel diplexer of this type, with strict specification in a proposed practical system [1], is given in the 80-GHz band with a 3-dB bandwidth of 800 MHz.

The proposed channel diplexer using a TE_{01} mode semicircular waveguide as the main guide has the same operational principle as the two-cavities ring-type diplexer, which is well known as a channel diplexer for the millimeter-wave channel multiplexing network [6]–[10]. A semicircular waveguide propagating the TE_{01} mode is chosen as the main guide by reason of the advantage in loss, relative to the rectangular guide, and in tight coupling characteristic, relative to the circular guide.

An experimental model in the 80-GHz band with a 3-dB bandwidth of 800 MHz was designed and fabricated. Good agreement has been achieved among design objectives, theoretical performance, and measured performance. Thus it is shown that a channel diplexer involving oversized waveguide can be practically designed using the proposed design technique.

II. Channel Diplexer Structure and Design

The structure of the proposed channel diplexer with a semicircular waveguide, employing two deformed ring cavities, is shown in Fig. 1. The channel diplexer is composed of two deformed ring cavities, three-directional couplers, a TE_{01} mode semicircular waveguide as the through-channel ports ① and ②, and a rectangular waveguide as the dropped-channel ports ③ and ④. For the design of the channel diplexer, the following four points are particularly taken into consideration.

1) No spurious mode should pass through cutoff in the main guide, and spurious resonant modes in the cavity must be avoided within the operating band.

2) Multislit E-plane coupled directional couplers, having a tight coupling characteristic, are applied because a 3-dB bandwidth needs more than 800 MHz to meet the extremely high clock frequency (400 MBd), and also excitation of spurious modes is sufficiently low.

3) The frequency spacing of the spurious resonance from the desired resonance is satisfied over 5 GHz at 80-GHz band because seven channels are connected in tandem with 740-MHz channel spacing.

4) Filter fabrication is devised so as not to increase the insertion loss caused by the imperfect electrical contact.

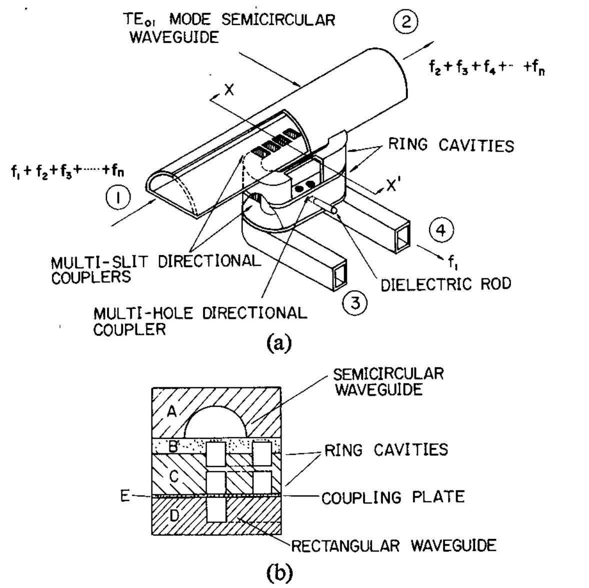


Fig. 1. Structure of a two-cavity ring-type channel diplexer with a semicircular waveguide. (a) Overall view. (b) $X - X'$ section.

With multihole directional couplers, less than 1 percent of bandwidth can be achieved if the frequency spacing between adjacent resonances is satisfied over 5 GHz. Thus newly devised multislit directional couplers are used instead of the multihole type [9]. The operating principle of the multislit coupler is well known [11]. The multislit E-plane coupled directional couplers are the upper and lower ones and the middle is a multihole coupler. This configuration is used because the amplitude coupling coefficient of the middle coupler is sufficiently smaller than that of the other couplers.

Coupling apertures are located along the center axis of the plane of bifurcation between the main waveguide and the resonant cavity. Only longitudinal magnetic field h_z, that will respond mainly to the TE_{01} mode, is coupled in the apertures. Because the aperture width is finite, the spurious modes will be excited by the aperture although the h_z component of these spurious modes vanishes at the center of the aperture. However, they are over 34 dB below the TE_{01} mode level for the aperture used in this design.

The diameter of the semicircular waveguide is chosen to fulfill the following requirements: 1) The TE_{02} mode should not pass through the cutoff because it is not absorbed by a helix waveguide. 2) The TE_{01} mode phase velocity is equal to that of the rectangular waveguide mode, so as not to produce reflection in the directional coupler.

Fig. 2 shows an equivalent circuit of the two-cavity ring-type channel diplexer. Coupling coefficients of each

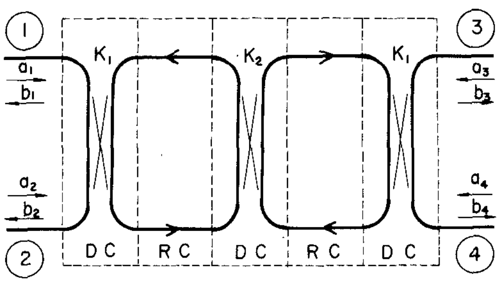


Fig 2. An equivalent circuit of the two-cavity ring-type channel diplexer. DC: Directional coupler. RC: Ring-type cavity.

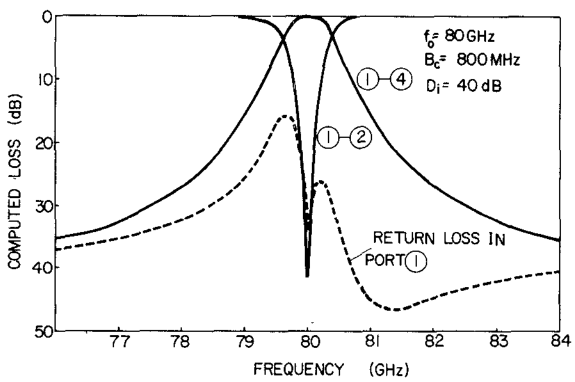


Fig. 3. Computed insertion loss and return loss of two-cavity channel diplexer centered at 80 GHz.

directional coupler in the vicinity of the center frequency under the condition of maximally flat response are given by the following equation:

$$
\begin{aligned}
\frac{f_0}{B_c} &= \sqrt{2}\,\pi N \frac{\sqrt{1 - K_1{}^2}}{K_1{}^2} \left(\frac{\lambda_g}{\lambda}\right)^2 \\
K_2 &= \frac{K_1{}^2}{2 - K_1{}^2} \\
\lambda_g &= \frac{\lambda}{\sqrt{1 - \left(\dfrac{\lambda}{\lambda_c}\right)^2}}
\end{aligned}
\tag{1}
$$

where

- λ wavelength in free space;
- λ_c cutoff wavelength;
- K_1 coupling coefficient between through/coupled channel and cavity;
- K_2 coupling coefficient between cavities;
- f_0 center frequency;
- B_c 3-dB bandwidth;
- N resonant index.

Resonant index N refers to the number of wave variations along the length of the ring cavity. Fig. 3 shows a theoretical performance of the two-cavity ring-type channel diplexer computed by the equation given in Appendix B, where it is assumed that f_0, B_c, and the directivity of each coupler D_i are 80 GHz, 800 MHz, and 40 dB (nearly equal to the average of measured value), respectively. Return loss (input VSWR) of the channel diplexer within the 3-dB bandwidth depends upon the directivity of each directional coupler [12].

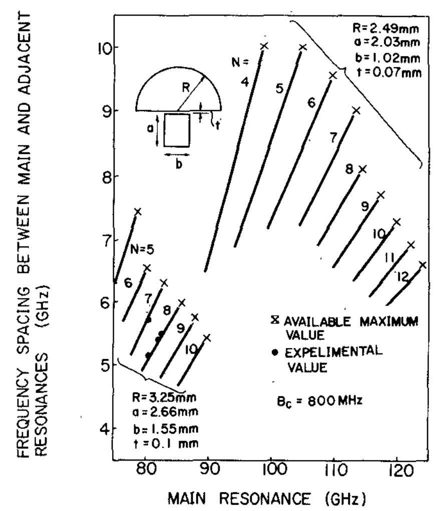


Fig. 4. Frequency spacing between main resonance and adjacent undesired resonance.

Frequency spacing B_f of the spurious resonance from the desired resonance is approximated as follows:

$$
B_f \doteqdot \frac{c}{\lambda_g}\left(\sqrt{\left(\frac{N+1}{N}\right)^2 + \left(\frac{\lambda_g}{\lambda_c}\right)^2} - \sqrt{1 + \left(\frac{\lambda_g}{\lambda_c}\right)^2}\right) \tag{2}
$$

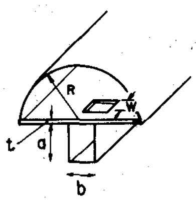
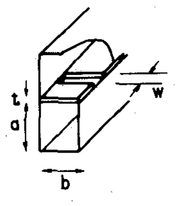
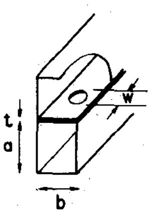
where c is the velocity of light. The 3-dB bandwidth B_c and the frequency spacing B_f depend on coupling coefficients K_1, K_2 and resonant index N. In order to obtain a broad 3-dB bandwidth, such as above 1 percent of the center frequency, and a wide frequency spacing, resonant index N must be made small and then the coupling per unit length has to be made as tight as possible. Fig. 4 shows an available frequency spacing between the frequencies of the adjacent modes.

Three kinds of coupling coefficients are shown in Table I. The small coupling theory is no longer accurate for a large aperture such as the slit. Therefore, the magnetic dipole moment of the semicircular-to-rectangular coupling aperture is estimated from the equation of the rectangular-to-rectangular E-plane coupled multislit coupler. The derivation of a coupling coefficient is shown in Appendix A. In order to hold the accuracy to the theoretical value, corrective constant ξ, established by measurements, is multiplied (see Fig. 5).

The presence of a coupling aperture on a cavity wall will perturb the longitudinal magnetic field and will alter the resonance frequency. Deviation δf of the resonance is given as the following approximation:

$$
\begin{aligned}
\delta f &= -\frac{\Delta\phi}{\phi_t}\left(\frac{\lambda}{\lambda_g{}^2}\right) c \\
\Delta\phi &= \phi_A + \phi_B + \phi_C \\
\phi_t &= 2\pi N \\
\phi_A &= \sin^{-1} K_1 + \sin^{-1} K_2 \\
\phi_B &= \frac{\pi t \lambda}{a^3} \sum_i W_i \left(\frac{\lambda}{\lambda_g}\right) \\
\phi_C &= -\frac{\pi^2 R}{3}\left(\frac{b}{R}\right)^2 \left[\frac{1}{2} - \frac{1}{5}\left(\frac{2\pi b}{\lambda_g}\right)^2\right]
\end{aligned}
\tag{3}
$$

TABLE I
COUPLING COEFFICIENTS

	$C_1 = \frac{X_{01}\lambda_g}{R^2 J_0(X_{01})} \cdot \frac{M}{\sqrt{2\pi a^3 b}} \cdot \delta_1 \eta_1 \xi$ $M = \frac{a^2 b}{2\pi\left[2+\left(\frac{4a}{\pi W}\right)^2\right]}$ $\delta_1 = e^{-\frac{\pi}{W}(0.02W+t)\sqrt{1-\left(\frac{2W}{\lambda}\right)^2}}$ $\eta_1 = \tan\left(\frac{\pi W}{\lambda}\right)/\left(\frac{\pi W}{\lambda}\right)$ $\lambda_g = \lambda/\sqrt{1-(\lambda/2a)^2}$, $\xi = 1.3$
	$C_2 = \frac{B}{2\sqrt{B^2+1}}\delta_1 \eta_1$ $B = \frac{\lambda g}{a\left[2+\left(\frac{4a}{\pi W}\right)^2\right]}$
	$C_3 = \frac{B}{2\sqrt{B^2+1}}\delta_3 \eta_3$ $B = \frac{\lambda g}{a\left(2+\frac{6a^2b}{\pi W^3}\right)}$ $\delta_3 = e^{-\frac{2\pi t}{1.71}\sqrt{1-\left(\frac{1.71W}{\lambda}\right)^2}}$ $\eta_3 = \tan\left(\frac{\pi}{2}\frac{1.71W}{\lambda}\right)/\left(\frac{\pi}{2}\frac{1.71W}{\lambda}\right)$

M : aperture magnetic polarizability
B : aperture susceptance
δ : coefficient representing wall thickness t effect
η : coefficient representing wave resonance effect in the aperture
ξ : corrective coefficient established by experiments

where

$a \times b$ rectangular waveguide section dimension ($a > b$);
ϕ_t total phase variation along the cavity at resonance;
$\Delta\phi$ phase perturbation;
W_i width of each slit aperture.

ϕ_A is caused by the reactance of the aperture [13]. With the existence of a slit aperture whose thickness is t, the waveguide looks as if its width widens from a to $a + t/2$, and accordingly, the phase velocity is varied. The total phase variation is given as ϕ_B. ϕ_C is caused because of the difference of the wavelength between the straight waveguide and the bent waveguide.

III. EXPERIMENTAL RESULTS

Channel diplexer specifications are shown as follows:

center frequency	$f_0 = 81.91$ GHz;
3-dB bandwidth	$B_c = 800$ MHz;
rectangular waveguide	$a \times b = 2.66 \times 1.55$ mm^2;
semicircular waveguide radius	$R = 3.25$ mm;
resonant index	$N = 8$;
frequency spacing	$B_f > 5$ GHz.

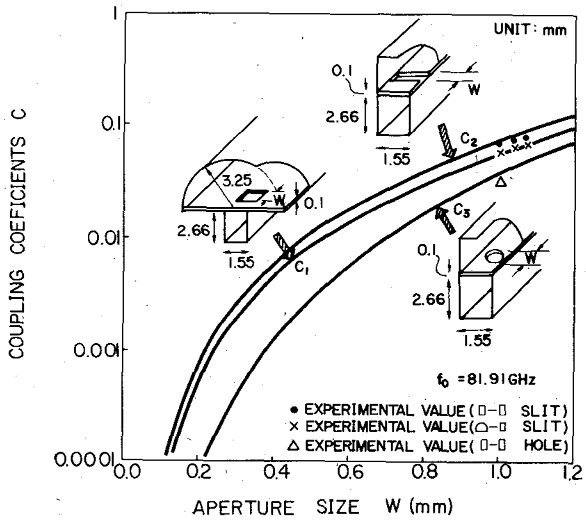
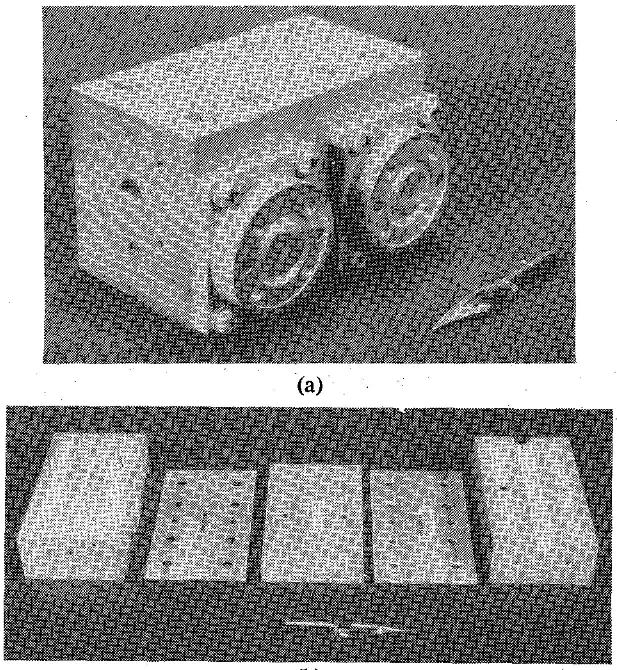


Fig. 5. Coupling coefficients of each directional coupler.



(a)

(b)

Fig. 6. Experimental model of the channel diplexer with a semicircular waveguide. (a) Overall view. (b) Fabrication method.

The experimental channel diplexer model is shown in Fig. 6. Its size is $68 \times 43 \times 47$ mm^3. The diplexer consists of four metal blocks, A through D, and a metal plate E, as shown in Fig. 1(b). The diplexer must be manufactured so as not to increase additional loss due to the imperfection of electrical contact among blocks and plate. Block B is the most difficult section to fabricate because it consists of 0.1-mm-thick coupling elements and a part of a cavity. They cannot be divided into two pieces because of the increasing loss due to the loose contact caused by insufficient pressure within the semicircular guide region. These blocks are fabricated by a milling machine. Coupling apertures are drilled by a discharge machine. Silver is plated on these surfaces. A dielectric rod is inserted into the cavities [see Fig. 1(a)] in order to cancel the backward wave caused by imperfect coupler directivity.

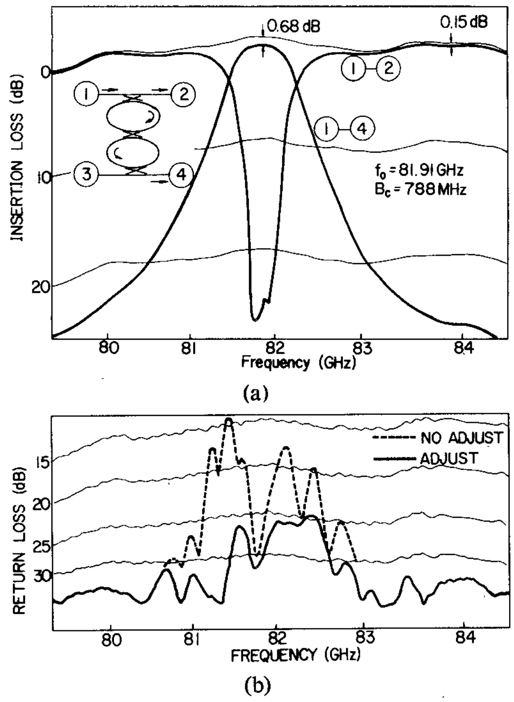


Fig. 7. Measured channel diplexer transmission characteristics. (a) Insertion loss. (b) Return loss.

TABLE II
MEASURED LOSS OF THE SEMICIRCULAR-TYPE DIPLEXER AND THE CONVENTIONAL RECTANGULAR ONE CENTERED AT 81.91 GHz WITH 800-MHz 3-dB BANDWIDTH

		Through Channel Loss (dB)	Channel dropping Loss (dB)	Return Loss (dB)
Semicircular Type	Computed Values	0.09	0.40	16
	Measured Values	0.15	0.68	14 (25*)
Conventional Rectangular Type	Computed Values	0.25	0.63	16
	Measured Values	0.32	0.99	23*

* Improved by dielectric rod which is inserted into the cavity and adjusted in order to cancel the reflection wave.

Fig. 7 shows representative characteristics of a diplexer. The channel-dropping loss L_B in ports ①–④ at a resonant frequency of 81.91 GHz and with 3-dB bandwidth of 788 MHz is 0.68 dB. The through-channel loss L_p in ports ①–② is 0.15 dB. These values show remarkable low loss compared with the conventional rectangular-type filter, as shown in Table II, where the theoretical losses are estimated based on the relative dc conductivities of silver. Input VSWR is less than 1.12. This shows that the return loss is improved from 14 to 25 dB by inserting the dielectric rod into cavities.

The theoretical center frequency deviation is calculated at 1.6 GHz by (3). In addition, dimensional tolerance of fabricated cavities must be taken into consideration, such that 1 μm corresponds to the deviation of 15 MHz. In this case, the resonant frequency of the cavities has been designed at 83.51 GHz. The deviation of the center frequency of four fabricated diplexers at the 80-GHz band were all within ± 30 MHz. The deviation of their 3-dB bandwidth were also within ± 30 MHz.

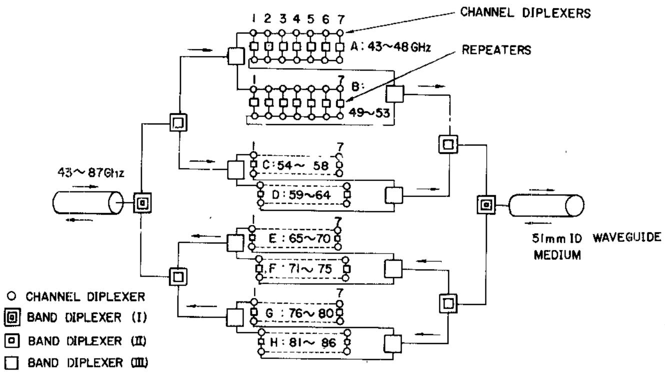


Fig. 8. Construction of a millimeter-wave channel multiplexing network using a 43–87-GHz frequency band.

Fig. 8 shows the construction of the millimeter-wave channel multiplexing network for the proposed millimeter-wave transmission system. This network consists of four kinds of diplexer: one band diplexer (I), two band diplexers (II), four band diplexers (III), and 56 channel diplexers. The last stage results from the use of eight multiplexers, from A to H, each with seven channels and a bandwidth of approximately 5 GHz. If the low-loss channel diplexer with a semicircular waveguide is applied to this network, the overall loss of its multiplexer decreases about 1.2 dB, as compared to the conventional rectangular-type filter. In addition, a mode transducer with 0.35-dB loss, which is otherwise required to be installed between the band diplexer using circular TE_{01} mode and the channel diplexer using rectangular TE_{10} mode, can be removed. Consequently, the total loss reduction is about 2 dB.

IV. CONCLUSION

The low-insertion-loss ring-type channel diplexer using a TE_{01} mode semicircular waveguide for the through channels has been designed and constructed in the 80-GHz band. Good agreement has been established between the theoretical calculation and the measured data. The channel-dropping loss over a 3-dB bandwidth of 788 MHz at the center frequency of 81.91 GHz is 0.68 dB. The through-channel loss is 0.15 dB, which is about half of that of the rectangular-type diplexer. The low through loss has an advantage for the connection of many channel diplexers in tandem, as is required in a guided millimeter-wave transmission system.

Thus the new two-cavity ring-type channel diplexer with a semicircular waveguide can be effectively used for guided millimeter-wave transmission system.

APPENDIX A

The coupling coefficient C between waveguides, depicted in Table I, is usually approximated, from small aperture theory, as the following equations:

$$C_1 = \frac{\pi}{\lambda_g} M h_z^{\frown} h_z^{\square} \quad \text{(A1)}$$

$$C_2 = \frac{\pi}{\lambda_g} M h_z^{\square 2} \quad \text{(A2)}$$

where

$$h_z^{\frown} = \frac{\lambda_g \chi_{01}}{\sqrt{2}\,\pi R}\frac{1}{\sqrt{\pi}}\frac{J_0(\chi_{01}\, r/R)}{R J_0(\chi_{01})}, \quad \text{for TE}_{01}\text{ semicircular waveguide}$$

$$h_z^{\square} = \frac{\lambda_g}{\sqrt{2a^3 b}}, \quad \text{for TE}_{10}\text{ rectangular waveguide}$$

$$\chi_{01} \risingdotseq 3.831706$$

where

- r distance from the center of the axis;
- M magnetic polarizability;
- λ_g wavelength in the guide.

However, in this case the aperture is not so small, compared with the wavelength where magnetic polarizability M cannot be determined accurately, and the preceding approximation is no longer applicable.

A large aperture E-plane slit coupling coefficient is known in [11] as follows:

$$C_2 = \frac{B}{2\sqrt{B^2+1}}$$

$$B = \frac{\lambda_g}{a[2 + (4a/\pi w)^2]}. \tag{A3}$$

Substituting (A3) into (A2), magnetic polarizability M is obtained as

$$M = Ba^3 b/2\pi\lambda_g. \tag{A4}$$

Then, from (A1) and (A4), the coupling coefficient between semicircular and rectangular waveguides is evaluated theoretically. The excitation level of a spurious mode coupled with the rectangular waveguide mode is also estimated by exchanging from the longitudinal magnetic field h_z of TE_{01} mode into that of the spurious mode in (A1). The calculation shows that the ratio of the coupling coefficients of the spurious modes (TE_{11}, TM_{01}, TE_{21}, and TM_{11}) and TE_{01} mode are 0, 0, 0.02 (−34 dB), 0, respectively.

Appendix B

The theoretical frequency response of the two-cavity ring-type channel diplexer is obtained by a matrix method. An equivalent circuit of the diplexer is shown in Fig. 2. An S matrix of the directional coupler is given as follows:

$$[b] = [S_{DC}][a]$$

$$[S_{DC}] = \begin{bmatrix} re^{-j\theta} & K'e^{-j\theta'} & -re^{-j\theta} & jK \\ K'e^{-j\theta'} & re^{-j\theta} & jK & -re^{-j\theta} \\ -re^{-j\theta} & jK & re^{-j\theta} & K'e^{-j\theta'} \\ jK & -re^{-j\theta} & K'e^{-j\theta'} & re^{-j\theta} \end{bmatrix} \tag{B1}$$

where

$$K'^2 = 1 - K^2 - 2r^2$$

$$\sin\theta' = -r^2/KK'$$

$$\cot\theta = \frac{-K'\sin\theta' - K}{K'\cos\theta'} \tag{B2}$$

and r and K are reflection and coupling coefficients, respectively. Similarly, an S matrix of the ring cavity is given as follows:

$$[S_{RC}] = \begin{bmatrix} 0 & 0 & e^{-j\beta l/2} & 0 \\ 0 & 0 & 0 & e^{-j\beta l/2} \\ e^{-j\beta l/2} & 0 & 0 & 0 \\ 0 & e^{-j\beta l/2} & 0 & 0 \end{bmatrix} \tag{B3}$$

where β is the wave number and l is the length of the ring cavity. Then, multiplying T matrixes transformed from the S matrixes of the directional coupler and ring cavity, a T matrix of the diplexer is obtained as follows:

$$[T_{\text{diplexer}}] = [T_{DC}][T_{RC}][T_{DC}][T_{RC}][T_{DC}] \tag{B4}$$

where $[T_{DC}]$ and $[T_{RC}]$ are transformed from each S matrix. Under the condition of $a_1 = 1, a_2 = a_3 = a_4 = 0$, the reflection, transmission and channel-dropping coefficients are equal to b_1, b_2, and b_4, respectively.

Acknowledgment

The authors wish to thank Dr. M. Shimba, M. Shinji, and Dr. M. Koyama of the Yokosuka Electrical Communication Laboratory for their invaluable advice. They also wish to thank K. Ohi of the Hitachi Electronics Company Ltd. for his endeavors in manufacturing diplexers, and T. Nunotani of the Shimada Physical and Chemical Industrial Company Ltd. for his valuable discussion.

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