

Overview of design method of millimeter-wave direct coupled resonator filters

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[Summary]

In recent years, the millimeter wave frequency band above 300 GHz has been developed for the practical application of next generation mobile communications. However, at a high frequency region such as a millimeter wave, several types of losses become pronounced, including dielectric losses of the substrate material. Thus, these losses limit the performance of analog frontend modules such as a millimeter-wave filter. This paper describes the design method for the type of direct coupled resonator as a typical waveguide filter. An example of design in the 300 GHz is compared with some previously reported microstrip filters focused on losses and filter performances. The limiting factors of filter performance and the necessity of precision machining for the waveguide filter processing are also discussed.

1 Introduction

The next generation information societies will consist of some key technologies: the Internet of Things (IoT), which enables big data collection and utilization, and the practical applications of virtual reality (VR) and augmented reality (AR). To realize these key technologies, high performance communication techniques, such as high speed, high capacity, low latency, and multiple simultaneous connections, will play an essential role^{1), 2)}. Current mobile communication technology, such as 5G, has begun utilizing not only microwave bands like the Sub 6 band but also millimeter-wave bands starting at 28 GHz, aiming to expand communication bandwidth. Furthermore, the next generation 6G is targeting even higher speeds and capacities and an increase in the number of connections by deploying in higher frequency millimeter-wave bands, such as the 300 GHz band. On the other hand, various applications of millimeter-wave sensing, which utilizes higher penetration compared to light and higher spatial resolution compared to microwaves, are also being explored^{3), 4)}. As such, the millimeter-wave band is actively being considered for usage. However, the design of passive devices in this frequency band requires a different approach than in the microwave band. This article outlines the design method of filters, those are the most important passive devices in the millimeter-wave range.

In the millimeter-wave range, circuit dimensions become comparable to the wavelength, and no inductors or capacitors are available with sufficient characteristics.

Hence, the filter design should be based on distributed element circuits consisting of the transmission line. Figure 1 shows typical transmission lines for microwave and millimeter-wave, those are utilized to constitute distributed element circuits. Filters consist of these transmission lines are compact such as less than several centimeters in the millimeter-wave region, whereas they are tens of centimeters in the microwave region. However, the decrease in size not only increases conductor losses, but also increases demands on manufacturing precision. The hollow waveguide shown in the upper part of Figure 1 has a low loss. However, due to its high cost of manufacture and large size in the microwave region, it has been limited to applications where high power handling is necessary. In the microwave region, a transmission line processed on a dielectric substrate, such as a microstrip, is widely used, as shown in the bottom part of Figure 1. These structures are easier to fabricate compared to hollow waveguides. However, the dielectric loss increases in a high frequency region such as the millimeter-wave.

Thus, appropriate filter geometries should be selected with consideration to both losses and ease of manufacture. Particularly in the millimeter-wave region, planar circuit filters on dielectric substrates are preferred when emphasizing integration density and ease of manufacture. In contrast, hollow waveguide filters are suitable when prioritizing low loss properties. In addition, the hollow waveguide filter can be directly connected to commonly used waveguides in the millimeter-wave band, thereby avoiding

waveguide type conversion losses. It is important in the front end of millimeter-wave measurement equipment that handles small level signals.

This article reviews the design procedure for bandpass filter (BPF) for millimeter-wave region and presents a design example of a 300 GHz band directly coupled resonator BPF based hollow waveguides. The designed BPF is compared with previously reported filters, and gains or losses of planar circuit filters on dielectric substrates versus the hollow waveguide filter, as well as the limiting factors of filter performance, are discussed.

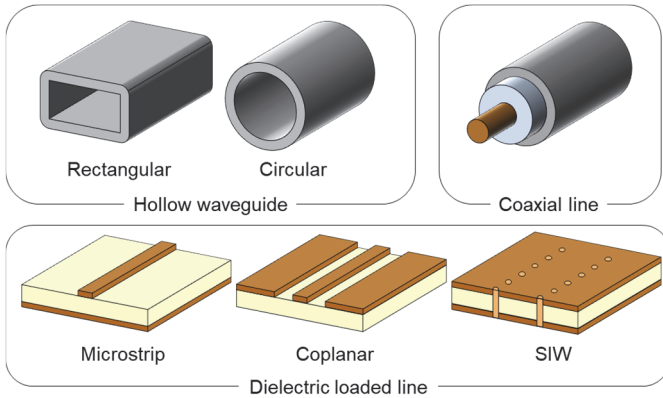


Figure 1 Various transmission lines used in the microwave and millimeter-wave bands

2 Design method of direct coupled resonator bandpass filter

BPF is one kind of passive circuit that allows only signals from a specific frequency band to pass through. For a method of its design, the insertion loss method is usually used, which was developed by G. Matthaei, L. Young, E. Jones, S. Cohn, and others in the 1950s. This method is based on the synthesis theory of networks and aims to achieve the desired frequency characteristics (insertion loss) by determining the circuit configuration and element values. In this procedure, first, a prototype low-pass filter (LPF) with normalized frequency and source impedance (admittance) is designed. Then, the circuit is transformed for ease of construction, followed by frequency scaling to convert it into a BPF. Inductors and capacitors deviate significantly from their ideal characteristics at the millimeter-wave range; hence, a discontinuity on a transmission line with reactance equivalent to these components is used to construct the filter. Electromagnetic field analysis is used to determine the

correspondence between the discontinuity dimensions on the line and the equivalent reactance. This relationship determines the structural dimensions that achieve the reactance designed by the insertion loss method. This chapter provides an overview of the above design procedure. For more details, please refer to references 5) to 8).

2.1 Frequency characteristics of bandpass filter

Typical specifications of frequency characteristics required for BPF include passband frequencies $f_{\text{pass,L}}$ and $f_{\text{pass,H}}$ with transmittance $|H_{\text{p,max}}|$, as well as stopband frequencies $f_{\text{stop,L}}$ and $f_{\text{stop,H}}$ with attenuation $|H_{\text{s,min}}|$. Figure 2 shows the typical relationship between BPF specifications and frequency characteristics. Filter design parameters such as cutoff frequencies f_1 , f_2 , and filter order N are defined to satisfy these specifications. A larger filter order provides a steep slope of the amplitude pass characteristic from near the cutoff frequency to the stopband.

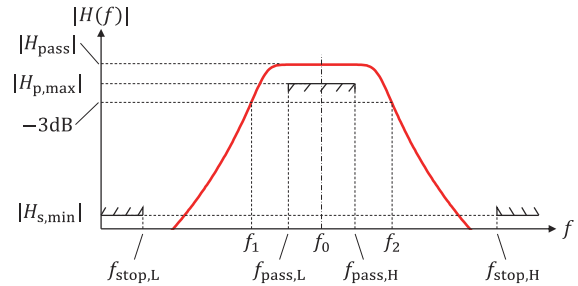


Figure 2 Typical BPF characteristics

2.2 Design of prototype LPF

A prototype LPF is a basic filter shape with the source impedance (or admittance) and the cutoff frequency normalized to unity. Figure 3 shows an equivalent ladder network of the prototype LPF, which consists of inductors and capacitors. Normalized element values (L'_i and C'_i) are represented as g_i . Figure 3 shows a sixth order design; the first reactance element is set to an inductor with a value of g_1 . $g_0 (= 1)$ and g_7 are the admittance of the source and the load, respectively. The same characteristics can also be composed with the other form of prototype LPF; the first reactance is set to a parallel capacitor (i.e., $g_1 = C'_1$), and every g_i represents impedances. These g parameters are set appropriately to obtain the required frequency response. Table 1 lists the typical frequency characteristics of some prototype filters, and Figure 4 shows the typical amplitude pass characteristics of each prototype LPF.

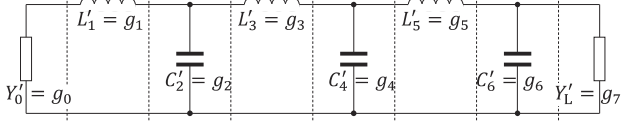


Figure 3 Prototype LPF equivalent circuit

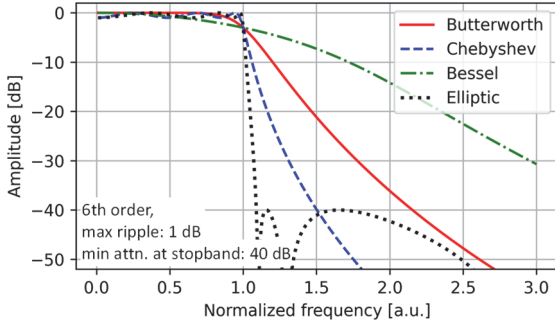


Figure 4 Amplitude pass characteristics of typical prototype LPFs

Table 1 Typical filter characteristics

Characteristic	Advantage	Disadvantage
Butterworth (maximally flat)	Highly flat response in passband	Not steep attenuation slope
Chebyshev (equiripple)	Steep attenuation slope	Passband is not flat, has ripple
Bessel (linear phase)	Linear phase change in passband, good temporal response	Very gradual attenuation slope, low frequency selectivity
Elliptical (equiripple in pass- and stop- band)	Attenuation slope can be steeply with placing attenuation pole properly.	Complicated structure due to interlaced coupling

In this paper, a Butterworth prototype LPF is used as an example. A Butterworth prototype LPF can be obtained by setting the element value g_i in Figure 3 to,

$$\begin{aligned} g_0 &= 1, \\ g_i &= 2 \sin \left[\frac{(2i-1)\pi}{2N} \right], \quad i = 1, 2, \dots, N, \\ g_{N+1} &= 1, \end{aligned} \quad (1)$$

where, N is the order of the filter, which corresponds to the number of ladder stages. The cutoff slope is determined to be $6N$ dB/oct., which corresponds to the filter order N . For example, in a sixth order Butterworth prototype LPF, the g parameter is shown:

$$g_0 = g_7 = 1.0000, \quad g_1 = g_6 = 0.5176, \\ g_2 = g_5 = 1.4142, \quad g_3 = g_4 = 1.9319.$$

The transfer function $H(s)$ of the Butterworth prototype LPF with such a g parameter set can be expressed as below,

$$H(s) = \begin{cases} \prod_{i=1}^{N/2} (s^2 + g_i s + 1)^{-1} & N: \text{even}, \\ \frac{1}{s+1} \prod_{i=1}^{(N-1)/2} (s^2 + g_i s + 1)^{-1} & N: \text{odd}. \end{cases} \quad (2)$$

2.3 Admittance Inverter

The prototype LPF shown in Figure 3 consists of both series and parallel connections of reactance elements. To facilitate implementation, use impedance inverters or admittance inverters to standardize either series or parallel connections. Here, we use an admittance inverter to convert the series inductor shown in Figure 5(a) to a parallel capacitor. An admittance inverter is a two-port network whose input admittance is J^2/Y_L when a load of admittance Y_L is connected. Here, J is called the admittance parameter (J parameter), and the J inverter's F matrix (transfer matrix) is shown,

$$F_J = \begin{bmatrix} 0 & 1 \\ -jJ & 0 \end{bmatrix}.$$

Here, the F matrix of the circuits shown in Figure 5(a) and (b) are below, respectively,

$$F_a = \begin{bmatrix} 1 & j\omega L \\ 0 & 1 \end{bmatrix}, \quad F_b = \begin{bmatrix} -\frac{J_2}{J_1} & -j\frac{\omega C_a}{J_1 J_2} \\ 0 & -\frac{J_1}{J_2} \end{bmatrix}. \quad (3)$$

If $F_a = F_b$, the following relation is obtained,

$$J_1 = -J_2, \quad L = -\frac{C_a}{J_1 J_2} = \frac{C_a}{J_1^2} = \frac{C_a}{J_2^2}.$$

Under the constraints imposed by this equation, circuits in Figure 5(a) and (b) can be made equivalent. Furthermore, as a general property of the circuit in Figure 5(b), the diagonal elements of F_b are not independent of each other, so F_b has two degrees of freedom: the (1,1) element and the (1,2) element. On the other hand, since the circuit has three parameters, J_1 , J_2 , and C_a , J_1 and J_2 can be determined such that $F_a = F_b$ for any C_a .

Therefore, the series inductors (L'_1 , L'_3 , L'_5) are shown in Figure 3 can be equivalently transformed into two J inverters and a parallel capacitor. Furthermore, all capacitances, including C'_2 , C'_4 , and C'_6 can be changed arbitrarily using the degree of freedom of J which was described above. As a result, the prototype LPF shown in Figure 3 can be converted into a circuit consisting only of parallel capacitors and J inverters, as shown in Figure 6. A J parameter $J'_{i,i+1}$ is given as follows for an arbitrary capacitance $C'_{a,i}$,

$$\begin{aligned} J'_{0,1} &= \sqrt{\frac{Y'_0 C'_{a,1}}{g_0 g_1}}, \\ J'_{i,i+1} &= \sqrt{\frac{C'_{a,i} C'_{a,i+1}}{g_i g_{i+1}}}, \quad i = 1, 2, \dots, N-1, \\ J'_{N,N+1} &= \sqrt{\frac{C'_{a,N} Y'_L}{g_N g_{N+1}}}. \end{aligned} \quad (4)$$

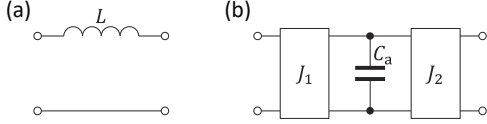


Figure 5 Circuit conversion by J inverter:
(a) series inductor, (b) capacitor between two J inverters

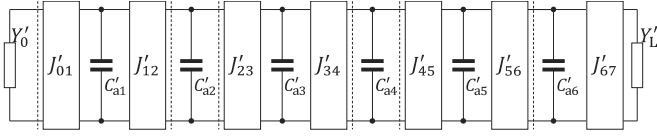


Figure 6 Equivalent circuit of prototype LPF with J inverter

2.4 Conversion from prototype LPF to BPF

Admittances are scaled from the prototype LPF shown in Figure 6 to the objective BPF. This scaling is performed by the conversion of the normalized angular frequency Ω in the prototype LPF and the angular frequency ω of the BPF as follows,

$$\Omega = \frac{\Omega_c}{\Delta} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) = \frac{1}{\Delta} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right). \quad (5)$$

Where ω_0 is the central angular frequency, which is expressed using cutoff angular frequencies $\omega_1 = 2\pi f_1$ and $\omega_2 = 2\pi f_2$ as follows,

$$\omega_0 = \sqrt{\omega_1 \omega_2}. \quad (6)$$

Then, the fractional bandwidth Δ is represented as below,

$$\Delta = \frac{\omega_2 - \omega_1}{\omega_0}. \quad (7)$$

According to this conversion, when $\omega > \omega_0$, it exhibits characteristics similar to the original LPF, and when $\omega < \omega_0$, it exhibits high pass characteristics in which the LPF is inverted at ω_0 . The cutoff angular frequency Ω_c of the prototype LPF corresponds to the cutoff angular frequency of objective BPF via the fractional bandwidth Δ .

The parallel capacitance C'_a in Figure 6 is converted as shown in Figure 7 using Eq. (5), then C_r and L_r are shown as follows,

$$\begin{aligned} j\Omega C'_a &= j\omega \frac{\Omega_c}{\omega_0 \Delta} C'_a + \frac{1}{j\omega} \frac{\omega_0 \Omega_c}{\Delta} C'_a, \\ C_r &= \frac{\Omega_c}{\omega_0 \Delta} C'_a, L_r = \frac{\Delta}{\omega_0 \Omega_c} \frac{1}{C'_a}. \end{aligned} \quad (8)$$

Then Eq. (4) is converted as follows,

$$\begin{aligned} J_{0,1} &= \sqrt{\frac{\omega_0 \Delta Y_0 C_{r,1}}{\Omega_c g_0 g_1}}, \\ J_{i,i+1} &= \frac{\omega_0 \Delta}{\Omega_c} \sqrt{\frac{C_{r,i} C_{r,i+1}}{g_i g_{i+1}}}, \quad i = 1, 2, \dots, N-1, \\ J_{N,N+1} &= \sqrt{\frac{\omega_0 \Delta C_{r,N} Y_L}{\Omega_c g_N g_{N+1}}}. \end{aligned} \quad (9)$$

Note that each capacitance $C_{r,i}$ can be set arbitrarily. This procedure provides a BPF consisting of the parallel LC resonant circuit and J inverter shown in Figure 8.

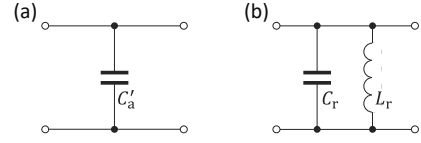


Figure 7 Circuit element conversion from prototype LPF to BPF:
(a) parallel C in prototype LPF, (b) parallel LC resonant circuit in BPF

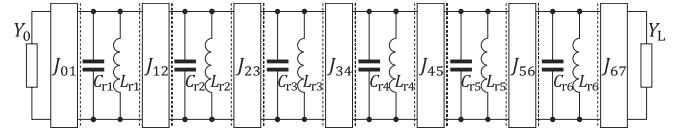


Figure 8 BPF consisting of parallel LC resonant circuit and J inverters

2.5 Conversion from lumped-element circuit to distributed-element circuit

The parallel LC resonant circuit in Figure 8 can be approximated using a cavity resonator constructed on a distributed element circuit. For this approximation, the susceptance of the cavity resonator must be equal to those of the parallel LC resonant circuit, and each first order derivative factor with respect to the frequency of the susceptance at the resonant frequency also must be equal to these simultaneously. The susceptance is defined as $B(\omega)$, and the slope parameter b respected to ω is defined as follows,

$$b = \frac{\omega_0}{2} \frac{dB(\omega)}{d\omega} \Big|_{\omega=\omega_0}. \quad (10)$$

Figure 9 shows a case of a transmission line used as a resonator. The susceptance of the parallel LC resonant circuit in Figure 9(a) is shown as follows,

$$B(\omega) = \omega C_r - \frac{1}{\omega L_r}, \quad \frac{dB(\omega)}{d\omega} = C_r \left(1 + \frac{\omega_0^2}{\omega^2} \right).$$

Since $B(\omega_0) = 0$ at the resonant angular frequency $\omega = \omega_0$, b was determined to,

$$b = \omega_0 C_r = \frac{1}{\omega_0 L_r}. \quad (11)$$

J parameters of each J inverters are determined by substituting Eq. (11) into Eq. (9), as follows,

$$\begin{aligned} J_{0,1} &= \sqrt{\frac{\Delta Y_0 b_1}{\Omega_c g_0 g_1}}, \\ J_{i,i+1} &= \frac{\Delta}{\Omega_c} \sqrt{\frac{b_i b_{i+1}}{g_i g_{i+1}}}, \quad i = 1, 2, \dots, N-1, \\ J_{N,N+1} &= \sqrt{\frac{\Delta b_N Y_L}{\Omega_c g_N g_{N+1}}}. \end{aligned} \quad (12)$$

On the other hand, for the transmission line with characteristic admittance Y_0 , phase constant β , length l , and open ended shown in Figure 9(b), the admittance $Y_{\text{line}}(\omega)$ seen from the input at angular frequency ω is shown as follows,

$$Y_{\text{line}}(\omega) = jY_0 \tan[\beta(\omega)l]. \quad (13)$$

Near the resonant angular frequency $\omega = \omega_0 + \delta\omega$ of the half-wavelength resonator is configured, assuming the phase constant β is linear around ω , the β is shown,

$$\beta(\omega_0)l = \pi, \quad \beta(\omega_0 + \delta\omega)l = \pi \left(1 + \frac{\delta\omega}{\omega_0}\right). \quad (14)$$

The first order derivation of the susceptance around ω_0 is shown,

$$\left. \frac{dB(\omega)}{d\omega} \right|_{\omega=\omega_0} = \frac{1}{j} \left. \frac{dY_{\text{line}}(\omega)}{d\omega} \right|_{\omega=\omega_0} \approx \frac{\pi Y_0}{\omega_0}. \quad (15)$$

From Eq. (10),

$$b \approx \frac{\pi}{2} Y_0, \quad (16)$$

is obtained. J parameters using resonator cavity are derived from Eq. (12) with $\Omega_c = 1$ and $Y_L = Y_0$, as follows,

$$\begin{aligned} \frac{J_{0,1}}{Y_0} &= \sqrt{\frac{\pi\Delta}{2} \frac{1}{g_0 g_1}}, \\ \frac{J_{i,i+1}}{Y_0} &= \frac{\pi\Delta}{2} \sqrt{\frac{1}{g_i g_{i+1}}}, \quad i = 1, 2, \dots, N-1, \\ \frac{J_{N,N+1}}{Y_0} &= \sqrt{\frac{\pi\Delta}{2} \frac{1}{g_N g_{N+1}}}. \end{aligned} \quad (17)$$

Furthermore, the J inverter can be converted to the π -type circuit shown in Figure 10. The J parameter normalized by characteristic admittance Y_0 is expressed as below⁵⁾,

$$\begin{aligned} \frac{J}{Y_0} &= \left| \tan \left[\frac{\phi}{2} + \tan^{-1} \left(\frac{B_p}{Y_0} \right) \right] \right|, \\ \phi &= -\tan^{-1} \left(2 \frac{B_s}{Y_0} + \frac{B_p}{Y_0} \right) - \tan^{-1} \left(\frac{B_p}{Y_0} \right). \end{aligned} \quad (18)$$

Here, B_p and B_s are the series susceptance and parallel susceptance of the π -type circuit, and ϕ represents the phase delay due to the π -type circuit. The line length l_i of the i -th resonator, when phase delay is considered, is represented by the equation below,

$$l_i = \frac{\lambda_{g0}}{2\pi} \left[\pi + \frac{1}{2} (\phi_i + \phi_{i+1}) \right], \quad i = 1, 2, \dots, N. \quad (19)$$

Here, λ_{g0} is the guide wavelength of the center frequency.

By conversion of the above procedure, Figure 8 is replaced with a $\lambda_g/2$ resonator and a π -type equivalent circuit, as shown in Figure 11.

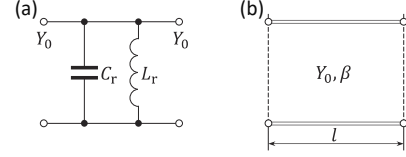


Figure 9 Conversion of a LC resonant circuit to a distributed-element circuit:

(a) parallel LC resonant circuit, (b) cavity resonator based on a waveguide with the electrical length π

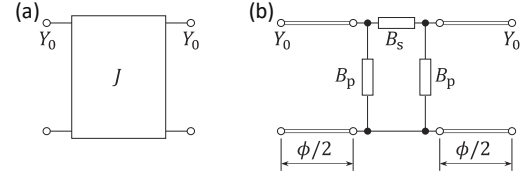


Figure 10 Example of equivalent circuit of J inverter:

(a) four-terminal model of J inverter, (b) symmetric π -type equivalent circuit containing a waveguide with the electrical length ϕ

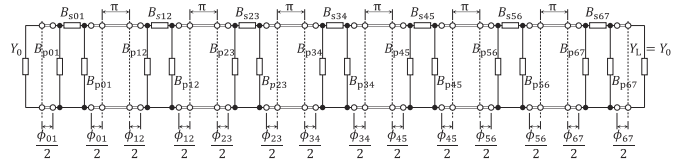


Figure 11 A BPF circuit constructed by $\lambda_g/2$ cavity resonators and π -type circuits

2.6 Discontinuous structure on waveguide and its electrical characteristics

The electrical characteristics of a discontinuous structure on a waveguide can be approximated by a π -type equivalent circuit. In this section, the inductive window in the rectangular waveguide is approximated by a π -type circuit. An inductive window in a rectangular waveguide has a conductor partition plate, which is inserted from either of the left and right sides, or both sides of a waveguide. With a sufficiently thin plate, its characteristics can be approximated by a parallel inductor. Figure 12 shows the procedure of calculating the parameters of a π -type equivalent circuit for an inductive window using an electromagnetic field (EMF) simulator (e.g., 3D EMF simulator based on a finite element method). As shown in Figure 12(a), a model is created for an inductive window with a width w and thickness t . The scattering matrix S is determined by an EMF simulator for each dimension parameter (w in here) of the discontinuous structure as in Figure 12(b). The susceptance of the π -type circuit can be determined from the S using the following equation (Figure 12(c)),

$$j \frac{B_p}{Y_0} = \frac{1 - S_{12} - S_{11}}{1 + S_{11} + S_{12}}, \quad (20)$$

$$j \frac{B_s}{Y_0} = \frac{2S_{12}}{(1 + S_{11})^2 - S_{12}^2}.$$

The width of the inductive window dimension (width: w) can be determined from the arbitrary equivalent susceptance.

With the above preparations, the dimensions of inductive windows w and cavity lengths l are determined as follows. J and ϕ can be found from Eq. (18) when w is given. Therefore, it is possible to find w such that J matches with $J_{i,i+1}$ designed from Eq. (17). Since J is a coupling coefficient between resonators, there always exists w that gives any J in the range of 0 to 1. Once the aperture width w is determined, the phase ϕ can also be found from Eq. (18), and the cavity length l of the $\lambda_g/2$ resonator can also be obtained from Eq. (19).

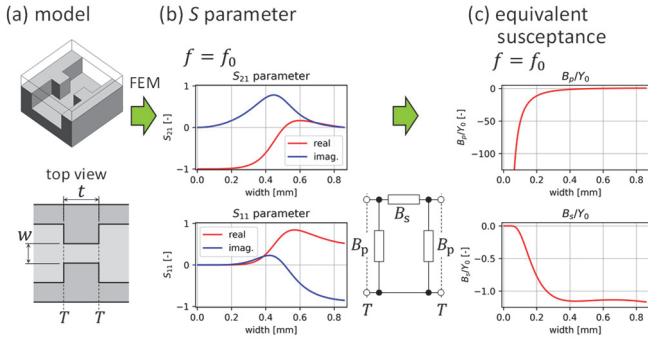


Figure 12 Equivalent parameter calculation process of a discontinuous structure using EMF simulator:

- (a) inductive window in rectangular waveguide (cut model),
- (b) S parameters obtained from EMF simulation,
- (c) equivalent susceptance in a π -type circuit

3 Design example of millimeter-wave filter and EMF simulation

3.1 Design of 300 GHz-band direct coupled resonator BPF

Table 2 shows the specifications and design parameters of the designed 300 GHz-band BPF. For the designed specifications, a center frequency and a passband width were set to 300 GHz and 10 GHz, respectively. The BPF was constructed in a WR-3.4 standard rectangular waveguide (operating frequency range is 220 to 330 GHz). A sixth order Butterworth filter was selected to satisfy the specifications, and cutoff frequencies were set to $f_1 = 293$ GHz and $f_2 = 307$ GHz. Following the design procedure in Chapter 2, direct coupled six $\lambda_g/2$ cavity resonators, which are

separated by seven inductive windows in the waveguide, were designed. The thickness of inductive windows was set to $t = 0.2$ mm. Table 3 shows the dimensions of each part determined through the design process. Figure 13 shows a cut-model of the designed BPF. The filter in which some resonators are directly coupled, as shown in Figure 13, hence it is called a direct coupled resonator filter.

Table 2 Design specifications and parameters

Required specifications			
Waveguide: WR-3.4 ($W \times H = 0.864 \times 0.432$ mm)			
$f_{\text{pass,L}}$: 295 GHz	$f_{\text{pass,H}}$: 305 GHz	$ H_{\text{p,max}} $	>-5 dB
$f_{\text{stop,L}}$: 280 GHz	$f_{\text{stop,H}}$: 320 GHz	$ H_{\text{s,min}} $	<-40 dB
Design parameters			
Filter type: sixth order Butterworth BPF			
f_1 : 293 GHz		f_2 : 307 GHz	
Window type: inductive window, $t = 0.2$ mm			

Table 3 Designed parameters and dimensions of Butterworth BPF using inductive window on WR-3.4 waveguide

	Width [mm]	Length [mm]	J / Y_0	ϕ [rad.]
Window 0-1/ Cavity 1	0.5094	0.3939	0.2466	1.8264
Window 1-2/ Cavity 2	0.3749	0.4992	0.0368	2.6052
Window 2-3/ Cavity 3	0.3268	0.5248	0.0190	2.7286
Window 3-4/ Cavity 4	0.3155	0.5248	0.0163	2.7524
Window 4-5/ Cavity 5	0.3268	0.4992	0.0190	2.7286
Window 5-6/ Cavity 6	0.3749	0.3939	0.0368	2.6052
Window 6-7	0.5094	—	0.2466	1.8264

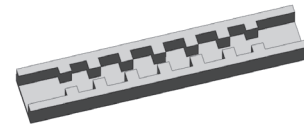


Figure 13 Picture of designed direct coupled resonator BPF (cut model)

3.2 Evaluation of designed filter characteristics

The designed filter was simulated by an EMF simulator (Femtet, Murata Software Co., Ltd.) to evaluate frequency characteristics. Figure 14 shows the reflectance and transmittance characteristics obtained by the simulation. Here, Figure 14(a) shows a model without loss, and (b) shows

results for some lossy models. In Figure 14(a), the 3 dB passband and the pass-band return loss were obtained to 293.8 to 306.2 GHz and <-30 dB, respectively. These results satisfy the design specification in Table 2. In Figure 14(b), to consider the effects of conductor losses, the following four models were compared: (i) with loss-less, (ii) with smooth copper (conductivity 6.0×10^7 S/m), (iii) with smooth aluminum alloy (conductivity 2.0×10^7 S/m) and (iv) with aluminum alloy with rough surface ($R_{\text{rms}} = 0.4$ μm). The maximum losses within the passband ($f_{\text{pass,L}}$ to $f_{\text{pass,H}}$) were 0.0 dB, 1.5 dB, 2.0 dB, and 3.3 dB, respectively. As the conductivity decreases or the surface becomes rougher, the loss increases. This result will be discussed again in Section 4.2. Note that even aluminum alloy with surface roughness $R_{\text{rms}} = 0.4$ μm satisfies the specifications in Table 2.

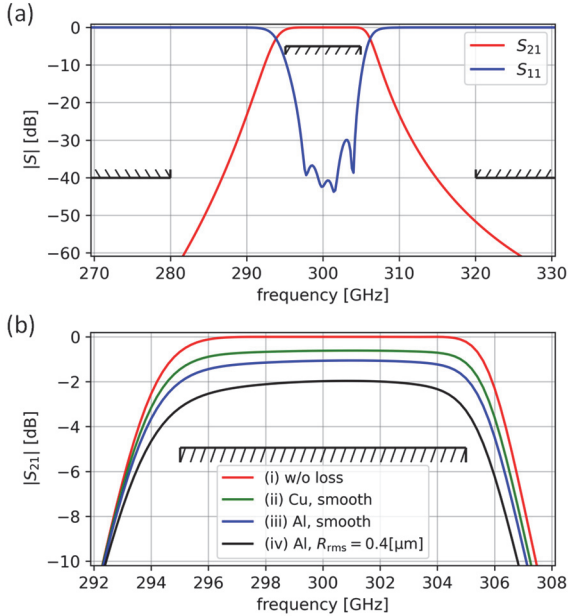


Figure 14 Simulated characteristics of the designed BPF:
(a) S_{11} and S_{21} under lossless condition,
(b) S_{21} with considering several varieties of losses; (i) loss-less, (ii) copper with smooth surface, (iii) Al alloy with smooth surface, (iv) Al alloy with rough surface ($R_{\text{rms}} = 0.4$ μm)

4 Notes for designing millimeter-wave filter

4.1 Filter characteristics variation by manufacturing process

The frequency characteristics of a filter composed of distributed-element circuits depend on each element's shape. Hence, manufacturing accuracy is required to be sufficiently appropriate for the objective frequency range. In this section,

the dependence of millimeter-wave filter characteristics on cavity dimensional error was evaluated using Monte Carlo simulation. Here, an 11th order Butterworth filter with a center frequency of 275 GHz and a passband width of 30 GHz was designed for both inductive and capacitive windows. Normal random numbers with a standard deviation of 2.5 μm were added as errors to the original design dimensions, and the transmittance characteristics $|S_{21}|$ were calculated for several pairs of dimensions. Figure 15 shows the simulation results; the dark line shows the design value model without any errors, and the light line shows the model with errors added. In the results of the inductive window Figure 15(a), (b), and capacitive window (c), deterioration is observed in the cutoff shoulder on the frequency characteristics due to dimensional errors. Figure 15(d) shows that sensitivity to a dimensional variation can decrease when the capacitive window is used, and the thickness is set to $t = \lambda_g/4$. The advantage of using a quarter wavelength thickness capacitive window for the cavity coupler was shown⁹⁾.

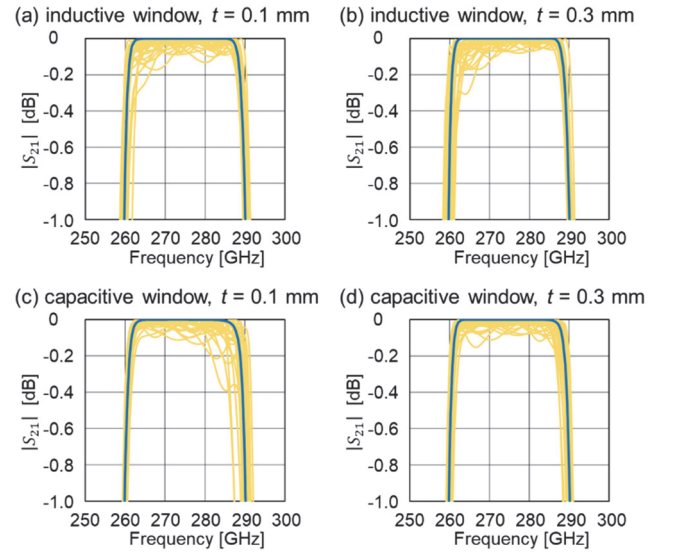


Figure 15 The filter characteristics depend on type of coupling window with dimensional variation, which simulates manufacturing accuracy

4.2 Filter performance limitations due to resonator losses

The loss in a resonator increases the insertion loss in the passband and limits the steep cutoff characteristics. The time-averaged ratio of the energy loss for a single cycle inside the resonator and the stored energy in the resonator is expressed as $2\pi/Q_u$. Here, Q_u is a dimensionless quantity

called the unloaded Q factor of the resonator. The smaller loss of the resonator becomes the larger Q_u and the steeper the resonance characteristics. The Q_u can be expressed by using Q_c due to conductor loss, Q_d by dielectric loss, and Q_r by radiation loss, as follows,

$$\frac{1}{Q_u} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_r} . \quad (21)$$

Note that magnetic loss is ignored here. For increasing Q_c and Q_d , it is necessary to use a low loss waveguide. A typical microstrip line (MSL) has more considerable radiation losses than hollow waveguides because of the electromagnetic field of an MSL existing in open space. Hence, shields are usually used to suppress radiation loss for increase Q_r . Assuming that the unloaded Q values of each resonator are the same, each loss can be compared by replacing to $s \rightarrow s + \delta$ in the transfer function $H(s)$ of prototype LPFs⁷⁾, where δ is a positive real number representing the loss, and $\delta = 1/(\Delta \cdot Q_u)$ for BPF. In particular, $|H(\delta)|$ represents the transmittance at the center frequency. Since the denominator of the transfer function is a polynomial for the variable s with positive coefficients, the transmittance decreases with smaller Q_u or narrower relative bandwidth. For example, the four models with different losses in Figure 14(b) verify the validity of this calculation. Table 4 shows the results of calculating the insertion loss at the center frequency from the transfer function. The relative bandwidth is $\Delta = 12.4/300$ from the 3 dB bandwidth 12.4 GHz of the lossless model in Figure 14(b). The unloaded Q for this verification was used based on the theoretical value of the TE₁₁₀ mode. As Table 4 shows, the theoretical values estimated from the transfer function agree well with the results of an EMF analysis.

Table 4 Comparison of insertion loss based on EMF analysis and theoretically from Eq. (21)

	EMF analysis	Theoretical value (Estimated)	
	Loss [dB]	Loss [dB]	unloaded Q
Copper (smooth)	0.64	0.52	1560
Aluminum alloy (smooth)	1.04	0.88	925
Aluminum alloy ($R_{\text{rms}}=0.4 \mu\text{m}$)	1.96	1.68	490

Table 5 shows the characteristics of previously reported filters using various waveguides and the filter designed in Section 3 for comparison. The bottom two rows in Table 5 show the unloaded Q and insertion loss converted to a center

frequency of 300 GHz, a passband width of 10 GHz, and a sixth filter order. Here, a Butterworth filter is assumed, and the unloaded Q is assumed to be inversely proportional to the square root of the center frequency. The loss calculation from the transfer function described above was used to calculate the insertion loss. These differences are due to the loss of each transmission line that constitutes the filter.

Table 5 Comparison of various BPF on several type of waveguides

Waveguide type	MSL	MSL	SIW (Si)	SIW (GaAs)	Hollow	Hollow (WR-3.4)
Center freq. [GHz]	89.6	152.4	270	328.5	286.6	300
Bandwidth [GHz]	11.7	10.5	5	46	16	10
Filter order	3	3	5	4	3	6
Attenuation @ f_0 [dB]	2.3	2.5	9.2	3.6	0.41	1.96*
References	(10)	(10)	(11)	(12)	(13)	—
Estimated Q @ 300 GHz	31.5	71.6	154	47.0	742	490
Estimated attenuation (sixth) [dB]	8.1	6.8	11.6	5.1	0.81	1.68

* Simulated in this paper. Al, $R_{\text{rms}}=0.4 \mu\text{m}$

To compare the losses in more detail, the cause-specific losses of typical waveguides, an MSL and a hollow waveguide, were calculated. Figure 16 shows losses of an MSL and a hollow in the 150 GHz and 300 GHz bands. The hollow waveguides were WR-6.5 and WR-3.4 as standard waveguides corresponding to each band, and the loss was determined for pure copper, aluminum alloy, and several surface roughness. The MSL was designed so that only the quasi-TEM mode propagated up to 170 GHz or 330 GHz, and its loss was simulated. The relative dielectric constant of the MSL substrate was set to $\epsilon_r = 2.0$ or 3.8, and the dielectric loss tangent was set 0.001 regardless of frequency. Each loss was calculated from the model in reference 6), considering the effect of surface roughness based on the Morgan-Hammastad model¹⁴⁾. Table 6 shows the attenuation constant of hollow waveguides, which is much lower than that of the MSL. Thus, an MSL has advantages such as a smaller footprint, but it should only be used for short distance connections where some loss is acceptable. In contrast, using the hollow waveguide, the attenuation constant of WR-3.4 remains at 20 dB/m with sufficiently low

surface roughness. Thus, it is effective when constructing a high-order filter for high-Q characteristics in the millimeter-wave region, or when suppressing insertion loss as much as possible is desired. To obtain even lower loss characteristics, surface finish processing such as lapping, which achieves a surface roughness of less than $R_{\text{rms}} = 0.1 \mu\text{m}$, is required. Therefore, it is desirable to realize a precision processing technology for a millimeter-wave filter that has both of the following specifications: low variation of manufacturing accuracy as shown in Section 4.1, low roughness of the surface as shown in Section 4.2.

Table 6 Comparison of attenuations of an MSL and a hollow waveguide

		Frequency	
		150 GHz	300 GHz
Attenuation [dB/m]	MSL	106	218
	Hollow waveguide (Cu, $R_{\text{rms}}=0.1 \mu\text{m}$)	6.2 (WR-6.5)	18.4 (WR-3.4)

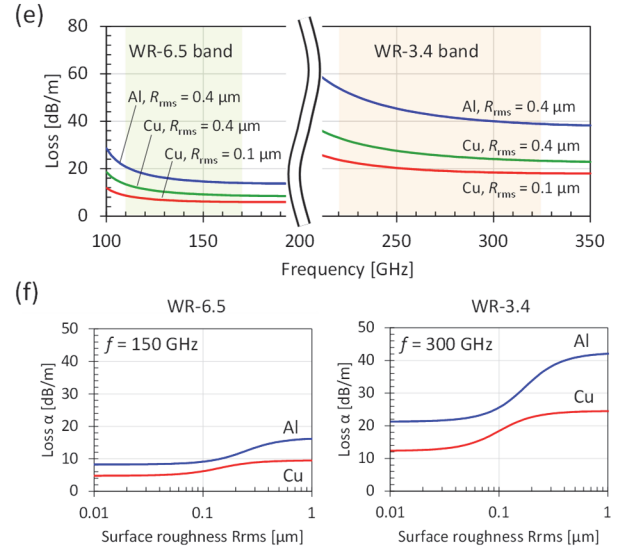
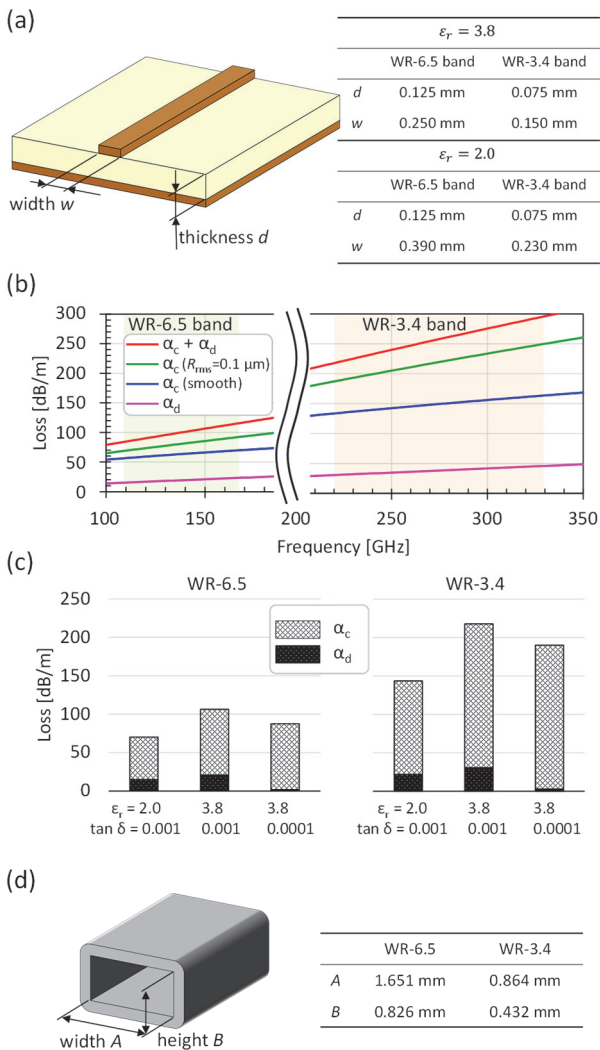


Figure 16 Propagation loss of an MSL and a rectangular hollow waveguide

- (a) MSL dimensions,
- (b) calculated loss results, α_c shows conductor loss, and α_d is dielectric loss, respectively,
- (c) loss comparison with various permittivity and dielectric loss tangent condition of MSL,
- (d) rectangular hollow waveguide dimensions,
- (e) calculated loss results due to various materials and surface roughness,
- (f) variation of loss due to waveguide surface roughness

5 Conclusion

This paper provided an overview of a filter design method and some notes for design in the millimeter-wave region. In next generation mobile communication, the use of frequency bands that extend to the sub-terahertz region is considered. In such a region, reducing losses of a passive device such as a filter is important. Construction on a low loss waveguide provides an effective way to improve the performance of a filter, and it is necessary to use them appropriately along with a filter on a planar circuit, which has better packaging density and ease of manufacturing.

In filter manufacturing, dimensional accuracy and surface roughness must be managed to achieve the required performance. However, the dimensional accuracy required in a high frequency of millimeter-wave region, such as the 300 GHz band, is generally close to the limit of processing technology. The application of precise manufacturing technology is anticipated to play a crucial role in consistently achieving the high-performance requirements of millimeter-wave band filters. Furthermore, a high-precision device

manufacturing technology for the millimeter-wave band is expected to advance by reducing variation due to manufacturing error and using a correction technique using compensating for the variation in the post process of trimming. These advanced technologies will help accelerate the practical application of the millimeter-wave band.

References

- 1) A. Shahraki, M. Abbasi, M. J. Piran, and A. Taherkordi, "A Comprehensive Survey on 6G Networks: Applications, Core Services, Enabling Technologies, and Future Challenges," arXiv:2101.12475, 2021.
- 2) W. Jiang, B. Han, M. A. Habibi, and H. D. Schotten, "The Road Towards 6G: A Comprehensive Survey," IEEE Open J. Commun. Soc., vol. 2, pp. 334-366, Feb. 2021.
- 3) M. Alizadeh, G. Shaker, J. C. M. Almeida, P. P. Morita, and S. S.- Naeini, "Remote Monitoring of Human Vital Signs Using mm-Wave FMCW Radar," IEEE Access, vol. 7, pp. 54958-54968, Apr. 2019.
- 4) S. Kharkovsky, and R. Zoughi, "Microwave and millimeter wave nondestructive testing and evaluation - Overview and recent advances," IEEE Instrum. Meas. Mag., vol. 61, iss. 5, pp. 2046-2060, May 2013.
- 5) G. L. Matthaei, E. M. T. Jones, and L. Young, "Microwave Filters Impedance-Matching Networks and Coupling Structures," Artech House, Norwood, MA, 1980.
- 6) D. M. Pozer, "Microwave Engineering," 4th Ed., John Wiley & Sons, Inc., New York, NY, 2011.
- 7) R. J. Cameron, C. M. Kudsia, and R. R. Mansour, "Microwave Filters for Communication Systems: Fundamentals, Design, and Applications," John Wiley & Sons, Inc., Hoboken, NJ, 2007.
- 8) Yoshio Kobayashi, Yoshinori Kogami, Yasuo Suzuki, "Microwave dielectric filters," IEICE, Tokyo, CORONA Publishing Co., Ltd, 2007 (in Japanese).
- 9) Shigenori Mattori, "Monte Carlo simulation on dependency on dimensional error of all-pole waveguide filters," 2023 IEICE General Conference, no.C-2-33, Saitama, Mar. 2023 (in Japanese).
- 10) Y. Chen and S.-W. Zhang, "Design of millimeter-wave bandpass filters based on improved asymmetric parallel-coupled microstrip line structure for direct detection system of CubeSat radiometer," Microw. Opt. Technol. Lett. 64(1), pp.23-29, Jan. 2021.
- 11) A. Krivovitca, U. Shah, O. Glubokov, and J. Oberhammer, "Micromachined Silicon-Core Substrate-Integrated Waveguides at 220–330 GHz," IEEE Trans. Microw. Theory Tech. 68(12), pp.5123-5131, Dec. 2020.
- 12) T.-Y. Chiu and C.-H. Li, "Low-Loss Low-Cost Substrate-Integrated Waveguide and Filter in GaAs IPD Technology for Terahertz Applications," IEEE Access 9, pp. 86346-86357, Jun. 2021.
- 13) H. Yang, Y. Dhayalan, X. Shang, M. J. Lancaster, B. Liu, H. Wang, M. Henry, and P. G. Huggard, "WR-3 Waveguide Bandpass Filters Fabricated Using High Precision CNC Machining and SU-8 Photoresist Technology," IEEE Trans. Terahertz Sci. Technol. 8(1), pp.100-107, Jan. 2018.
- 14) B. C. Wadell, "Transmission Line Design Handbook," Artech House, Boston, MA, 1991.

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