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TO COMPACT WAVEGUIDE DEVICES BY DIELECTRIC AND FERRITE LAYERS

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Abstract—In this paper, a method is proposed to compact waveguide devices at a desired frequency. In this method, a previously designed hollow waveguide is filled with several dielectric and ferrite layers alternately so that the characteristic impedance of the waveguide is not changed. First, the permittivity and permeability of a fictitiously mixed material is obtained. Then, the required permittivity and permeability of dielectric and ferrite layers are obtained at desired frequency. The usefulness of the proposed method is verified using some theoretical and simulation examples.

1. INTRODUCTION

In many applications we would like to compact RF and microwave circuits. On the other hand, many RF and microwave circuits contain one or some single transmission lines of specified length to work at a desired frequency. Therefore, one possible way to compact the circuits is to compact (to reduce the length of) the transmission lines. Some efforts have been made to compact the transmission lines such as using DGS [1], EBG [2–4], high impedance and meandering lines [5], fractal lines [6], stepped stubs [7] and nonuniform transmission lines [8]. Of course, most efforts have been made to compact microstrip devices not waveguide ones. In addition to use slow wave structures [3, 4], one way to compact waveguide devices is to fill the hollow waveguide with dielectric layers [9]. However, filling the waveguides with dielectric layers affects the performance of the device, and we have to redesign it considering the presence of dielectric layers. In this paper, we propose a new method to compact the waveguide devices. In this method a

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previously designed hollow waveguide is filled with several dielectric and ferrite layers alternately so that the characteristic impedance of the waveguide is not changed. In fact, the resulted waveguide operates as a matched slow wave structure. So in this method, it is not necessary to redesign waveguide devices considering the presence of dielectric and ferrite layers. First, the permittivity and permeability of a fictitiously mixed material is obtained. Then, the required permittivity and permeability of dielectric and ferrite layers are obtained at desired frequency. The usefulness of the proposed method is verified using some theoretical and simulated examples.

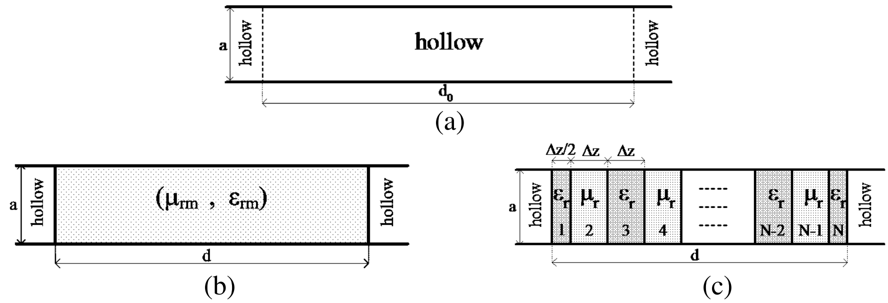

Figure 1. (a) The top view of a hollow waveguide of length d_0. (b) The top view of a waveguide of length d filled by a fictitious mixed material (MMW). (c) The top view of a waveguide of length d filled by N dielectric and ferrite layers alternately (DFW).

2. TO COMPACT HOLLOW WAVEGUIDES

In this section, the idea to compact the hollow waveguides by filling them with suitable materials is presented. Fig. 1(a) shows the top view of a hollow waveguide of length d_0, which we wish to compact. Fig. 1(b) shows the top view of a waveguide of length d smaller than d_0, which has been filled with a fictitiously mixed material of permittivity ε_{rm} and permeability μ_{rm}, calling it Mixed Material Waveguide (MMW). Also, Fig. 1(c) shows the top view of a waveguide of length d smaller than d_0, which has been filled with N dielectric ε_r and ferrite μ_r layers, alternately, calling it Dielectric Ferrite Waveguide (DFW). It is proven in the next section that the performance of the DFWs can be an approximation of that of MMWs. The cross section of all three waveguides have dimensions a and b. Also, it is obvious that only the dominant mode TE_{10} exists in all three waveguides. The thickness Δz

in Fig. 1(b) is given by

$$\Delta z = \frac{d}{N-1} \tag{1}$$

We would like the performance of all three waveguides to be identical at desired frequency f_0 while the length of material filled waveguides is smaller than that of the hollow waveguide by a compactness ratio $K:1$, i.e.,

$$d = \frac{d_h}{K} < d_h \tag{2}$$

The chain parameter matrix of the hollow waveguide is given by

$$\mathbf{T}_h = \begin{bmatrix} A_h & B_h \\ C_h & D_h \end{bmatrix} = \begin{bmatrix} \cos(k_h d_h) & jZ_h \sin(k_h d_h) \\ jZ_h^{-1} \sin(k_h d_h) & \cos(k_h d_h) \end{bmatrix} \tag{3}$$

where k_h and Z_h are the propagation coefficient and the characteristic impedance of the hollow waveguide, respectively, given by

$$k_h = k_0\sqrt{1-(f_c/f)^2} \tag{4}$$

$$Z_h = \frac{\eta_0}{\sqrt{1-(f_c/f)^2}} \tag{5}$$

in which $k = 2\pi f/c$ (c is the velocity of the light) is the wave number in the free space.

2.1. Design of MMWs

The chain parameter matrix of the MMW is given by

$$\mathbf{T}_m = \begin{bmatrix} A_m & B_m \\ C_m & D_m \end{bmatrix} = \begin{bmatrix} \cos(k_m d) & jZ_m \sin(k_m d) \\ jZ_m^{-1} \sin(k_m d) & \cos(k_m d) \end{bmatrix} \tag{6}$$

where k_m and Z_m are the propagation coefficient and the characteristic impedance of the MMW, respectively, given by

$$k_m = k_0\sqrt{\mu_{rm}\varepsilon_{rm}-(f_c/f)^2} \tag{7}$$

$$Z_m = \frac{\eta_0\mu_{rm}}{\sqrt{\mu_{rm}\varepsilon_{rm}-(f_c/f)^2}} \tag{8}$$

To make the elements of the chain parameter matrices of the hollow waveguide (3) equal to those of the MMW (6) at desired frequency f_0, two following relations have to be kept by equating the phase and the characteristic impedance of the MMW to those of the hollow waveguide.

$$K = \frac{k_m}{k_h} = \frac{\sqrt{\mu_{rm}\varepsilon_{rm}-(f_c/f_0)^2}}{\sqrt{1-(f_c/f_0)^2}} \tag{9}$$

$$\mu_{rm}^2\left(1-(f_c/f_0)^2\right) - \varepsilon_{rm}\mu_{rm} + (f_c/f_0)^2 = 0 \tag{10}$$

Consequently, the required mixed material has to have the following permittivity and permeability obtained from (9) and (10).

$$\mu_{rm} = K \tag{11}$$

$$\varepsilon_{rm} = K - \frac{K^2-1}{K}(f_c/f_0)^2 \tag{12}$$

It is seen that the required permittivity is dependent on desired frequency. So, the introduced compacting method is useful for narrowband waveguide devices. Moreover, the rate of variation of transmitting phase with respect to frequency for MMW is less than that for the hollow waveguide. The following relation can be obtained.

$$K' \triangleq \left(\frac{d\Phi_m}{df}\Big/\frac{d\Phi_h}{df}\right)\Big|_{f=f_0} = \left(1+\frac{(f_c/f_0)^2}{K^2\left(1-(f_c/f_0)^2\right)}\right)\left(1+\frac{(f_c/f_0)^2}{1-(f_c/f_0)^2}\right)^{-1} < 1 \tag{13}$$

where $\Phi_m = -k_m d$ and $\Phi_h = -k_h d_h$ are the transmitting phase of MMW and hollow waveguide, respectively.

It is worth to mention that another solution of (9) and (10) is a ferrite material (not a mixed material) with permittivity $\varepsilon_{rm} = 1$ and the following permeability

$$\mu_{rm} = K = \frac{(f_c/f_0)^2}{1-(f_c/f_0)^2} \tag{14}$$

This special case is very sensitive to f and can be used only when $f_c < f_0 < \sqrt{2} f_c$ to have K and μ_{rm} greater than one.

2.2. Design of DFWs

The construction of desired mixed materials is not practical usually. So, we approximate the mixed material by N alternating dielectric and ferrite layers as shown in Fig. 1(c). The chain parameter matrix of the DFW can be written as follows

$$\mathbf{T} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} = (\mathbf{T}_d \mathbf{T}_f \mathbf{T}_d)^{(N-1)/2} \tag{15}$$

in which $\mathbf{T}_d$ and $\mathbf{T}_f$ are the chain parameter matrices of the dielectric and ferrite layers, respectively as follows

$$\mathbf{T}_d = \begin{bmatrix} \cos(k_d \Delta z/2) & jZ_d \sin(k_d \Delta z/2) \\ jZ_d^{-1} \sin(k_d \Delta z/2) & \cos(k_d \Delta z/2) \end{bmatrix} \tag{16}$$

$$\mathbf{T}_f = \begin{bmatrix} \cos(k_f \Delta z) & jZ_f \sin(k_f \Delta z) \\ jZ_f^{-1} \sin(k_f \Delta z) & \cos(k_f \Delta z) \end{bmatrix} \tag{17}$$

In (16) and (17), the following propagation coefficients and characteristic impedances exist.

$$k_d = k_0\sqrt{\varepsilon_r - (f_c/f)^2} \tag{18}$$

$$k_f = k_0\sqrt{\mu_r - (f_c/f)^2} \tag{19}$$

$$Z_d = \frac{\eta_0}{\sqrt{\varepsilon_r - (f_c/f)^2}} \tag{20}$$

$$Z_f = \frac{\eta_0\mu_r}{\sqrt{\mu_r - (f_c/f)^2}} \tag{21}$$

The chain parameter matrices of dielectric and ferrite layers assuming $\Delta z \ll \lambda_0$, where λ_0 is the wavelength in free space at frequency f, can be approximated by

$$\mathbf{T}_d \cong \begin{bmatrix} 1 & j\eta_0 k_0 \Delta z/2 \\ j\eta_0^{-1} k_0 \left(\varepsilon_r - (f_c/f)^2\right) \Delta z/2 & 1 \end{bmatrix} \tag{22}$$

$$\mathbf{T}_f \cong \begin{bmatrix} 1 & j\eta_0 \mu_r k_0 \Delta z \\ j\eta_0^{-1} \mu_r^{-1} \left(\mu_r - (f_c/f)^2\right) k_0 \Delta z & 1 \end{bmatrix} \tag{23}$$

Therefore, the chain parameter matrix of a DFW with thickness $2\Delta z \ll \lambda_0$ can be approximated by

$$\begin{aligned} \mathbf{T}_{df} &= \mathbf{T}_d \mathbf{T}_f \mathbf{T}_d \\ &\cong \begin{bmatrix} 1 & j\eta_0(\mu_r+1) k_0 \Delta z \\ j\eta_0^{-1}\left(\mu_r^{-1}\left(\mu_r - (f_c/f)^2\right) + \left(\varepsilon_r - (f_c/f)^2\right)\right) k_0 \Delta z & 1 \end{bmatrix} \end{aligned} \tag{24}$$

On the other hand, the chain parameter matrix of the MMW (6) with thickness $2\Delta z \ll \lambda_0$ can be approximated by

$$\mathbf{T}_m \cong \begin{bmatrix} 1 & j2\eta_0 \mu_{rm} k_0 \Delta z \\ j2\eta_0^{-1} \mu_{rm}^{-1} \left(\mu_{rm}\varepsilon_{rm} - (f_c/f)^2\right) k_0 \Delta z & 1 \end{bmatrix} \tag{25}$$

Comparing (24) with (25), the following relations are obtained between the permittivity and permeability of the mixed material with those of dielectric and ferrite layers.

$$\mu_{rm} = \frac{\mu_r + 1}{2} \tag{26}$$

$$\varepsilon_{rm} = \frac{\varepsilon_r + 1}{2} - \frac{(\mu_r - 1)^2}{2\mu_r(\mu_r + 1)}(f_c/f)^2 \tag{27}$$

It is seen that the equivalent permittivity varies with respect to frequency. It is the second reason that the introduced compacting method is useful for narrowband waveguide devices.

Therefore, to compact hollow waveguides we first use (11) and (12) to obtain the parameters of the required mixed material. Then, the required permittivity and permeability of dielectric and ferrite layers can be obtained from (26) and (27) at desired frequency.

3. EXAMPLES AND RESULTS

In this section, the validation of the proposed compacting method is verified. Consider a rectangular waveguide of dimensions $a = 0.9$ inch and $b = 0.4$ inch (WR-90). Figs. 2–4 compare the $ABCD$ parameters of

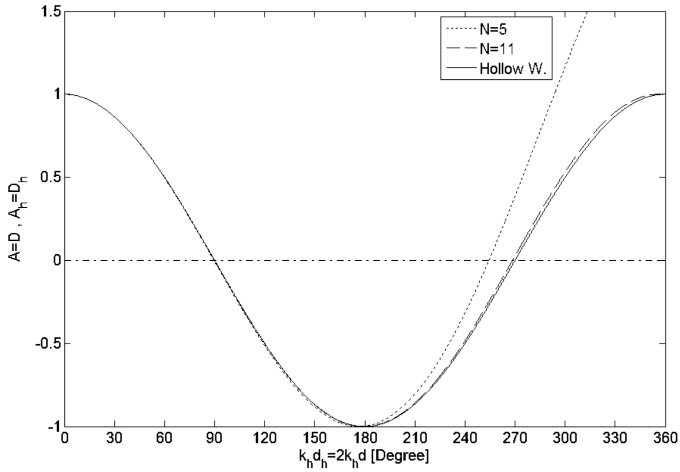

Figure 2. The parameters A and D of a DFW and a hollow waveguide for $K = 2$ and $f_0 = 10\,\mathrm{GHz}$ ($\mu_r = 3$ and $\varepsilon_r = 1.85$).

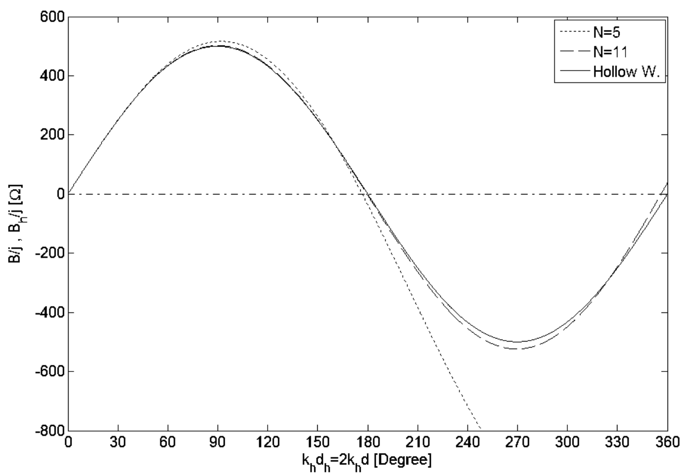

Figure 3. The parameter B of a DFW and a hollow waveguide for $K = 2$ and $f_0 = 10\,\mathrm{GHz}$ ($\mu_r = 3$ and $\varepsilon_r = 1.85$).

a DFW with those of a hollow waveguide assuming compactness ratio $K = 2$ and desired frequency $f_0 = 10\,\text{GHz}$ considering $N = 5$ and 11 layers. The required electrical parameters have been obtained as ($\mu_{rm} = 2$, $\varepsilon_{rm} = 1.35$) and ($\mu_r = 3$, $\varepsilon_r = 1.85$). Also, Fig. 5 illustrates

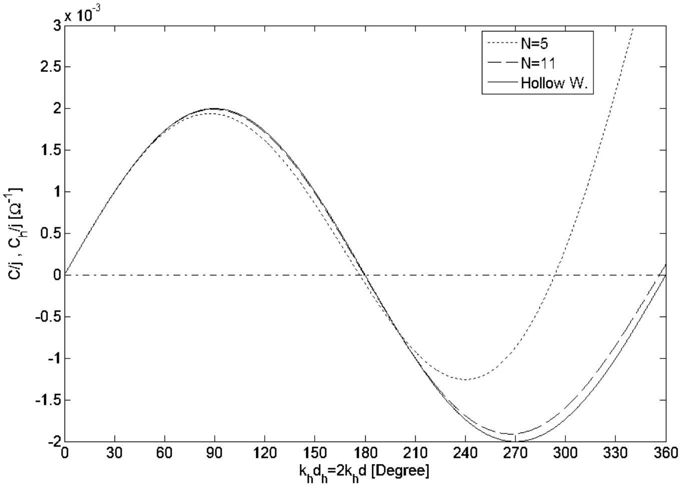

Figure 4. The parameter C of a DFW and a hollow waveguide for $K = 2$ and $f_0 = 10\,\text{GHz}$ ($\mu_r = 3$ and $\varepsilon_r = 1.85$).

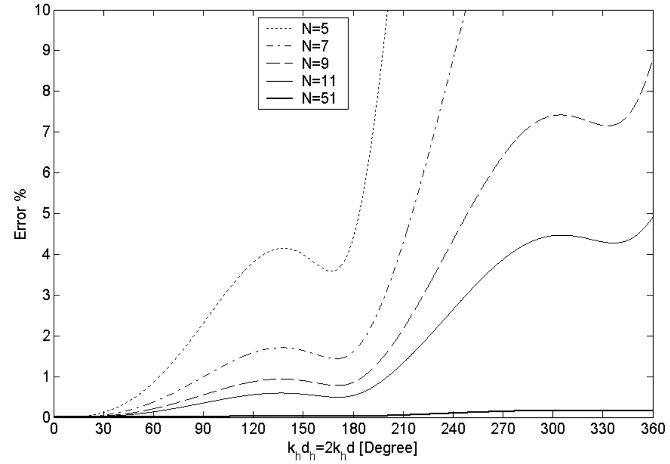

Figure 5. The error between $ABCD$ parameters of DFW and hollow waveguide for $K = 2$ and $f_0 = 10\,\text{GHz}$ ($\mu_r = 3$ and $\varepsilon_r = 1.85$).

the following defined error for $N = 5, 7, 9, 11$ and 51.

$$\text{Error} = \sqrt{\frac{1}{4}\left(|A-A_h|^2 + Z_h^{-2}\,|B-B_h|^2 + Z_h^2\,|C-C_h|^2 + |D-D_h|^2\right)} \quad (28)$$

Moreover, Fig. 6 illustrates the defined error for $f_0 = 9, 10, 11$ and 12 GHz considering $N = 11$. It is seen from Figs. 2–6 that the $ABCD$ parameters of DFW approach those of hollow waveguide more and more as the electrical length of hollow waveguide decreases; the number of layers N is increased, or the desired frequency decreases. In fact, as the electrical length of dielectric and ferrite layers is decreased the accuracy of approximation MMWs by DFWs is increased.

The chain parameter matrix of a DFW can be used to find its S parameters, as follows

$$S_{11} = \frac{(A - Z_hC)Z_h + (B - Z_hD)}{(A + Z_hC)Z_h + (B + Z_hD)} \quad (29)$$

$$S_{21} = S_{12} = \frac{2Z_h}{(A + Z_hC)Z_h + (B + Z_hD)} \quad (30)$$

$$S_{22} = \frac{(-A - Z_hC)Z_h + (B + Z_hD)}{(A + Z_hC)Z_h + (B + Z_hD)} \quad (31)$$

Figs. 7 and 8 illustrate the parameters S_{11} and S_{21} of DFW with those of a hollow waveguide assuming $k_hd_h = 90°$, $K = 2$ and $f_0 = 10$ GHz versus frequency considering $N = 5, 7, 9$ and 11 layers. An excellent

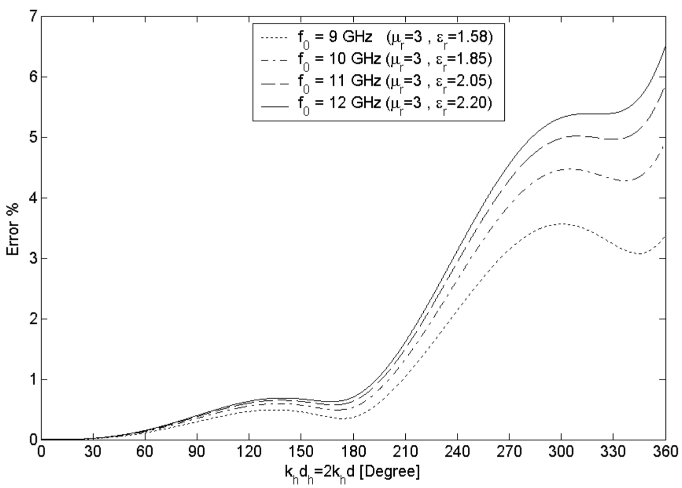

Figure 6. The error between $ABCD$ parameters of DFW and hollow waveguide for $K = 2$ and $N = 11$.

agreement between the parameters of DFW and hollow waveguide is seen at desired frequency, which is increased as the number of layers N is increased. Figs. 9 and 10 illustrate the amplitude of S_{21} of DFW and the difference between the phase of S_{21} of DFW and that of the hollow waveguide, respectively, versus frequency for $f_0 = 8, 9, 10, 11$ and $12\,\mathrm{GHz}$ considering $N = 11$. Also, Fig. 11

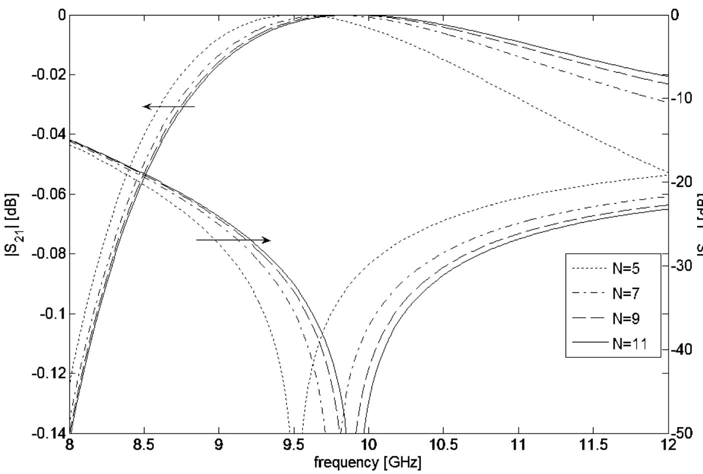

Figure 7. The amplitude of S_{11} and S_{21} of DFW for $k_h d_h = 90°$, $K = 2$ and $f_0 = 10\,\mathrm{GHz}$ ($\mu_r = 3$ and $\varepsilon_r = 1.85$).

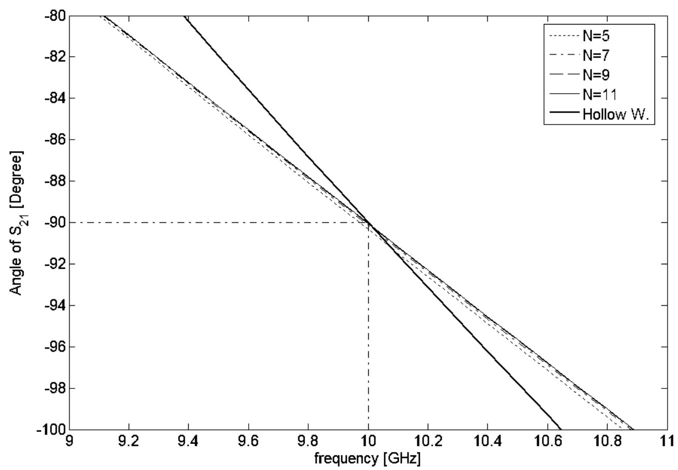

Figure 8. The phase of S_{21} of DFW and hollow waveguide for $k_h d_h = 90°$, $K = 2$ and $f_0 = 10\,\mathrm{GHz}$ ($\mu_r = 3$ and $\varepsilon_r = 1.85$).

illustrates the difference between the phase of S_{21} of DFW and that of the hollow waveguide versus frequency for $f_0 = 10\,\mathrm{GHz}$ considering $N = 11$ and $K = 5/4, 4/3, 3/2$ and 2. It is seen from Figs. 9–11 that as the desired frequency increases, or the compactness ratio is decreased the S parameters of DFW approach those of hollow waveguide in a wider bandwidth.

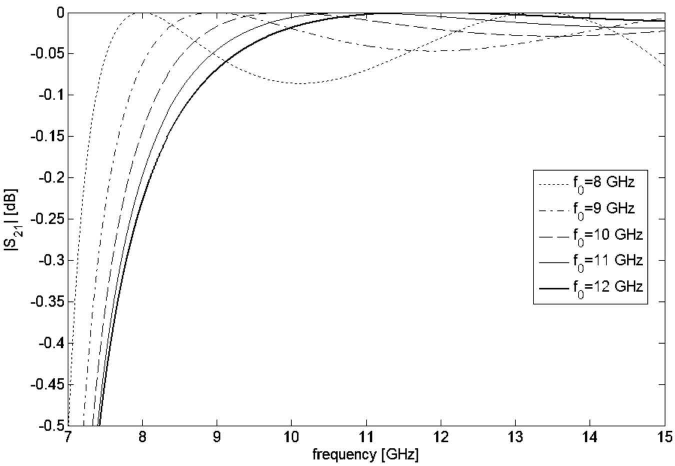

Figure 9. The amplitude of S_{21} of DFW for $k_h d_h = 90°$, $K = 2$ and $N = 11$.

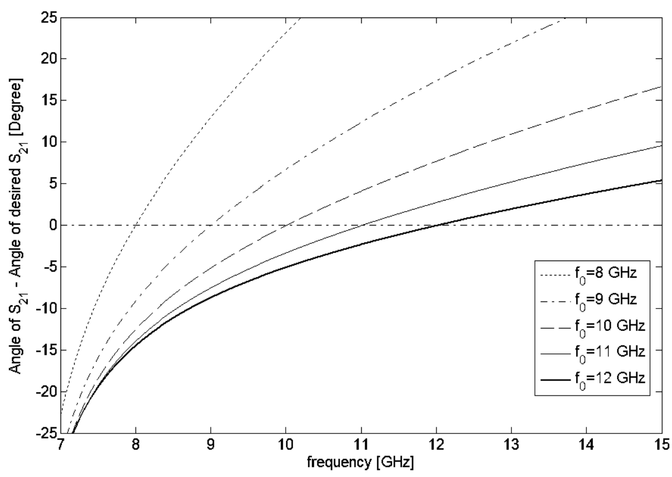

Figure 10. The difference between the phase of S_{21} of DFW and that of hollow waveguide for $k_h d_h = 90°$, $K = 2$ and $N = 11$.

Finally, consider an air-filled waveguide bandpass filter with inductive diaphragms designed in [9], whose characteristics are 3-order chebyshev type, center frequency 10 GHz, relative bandwidth 9 percent and equal ripples 0.5 dB. Fig. 12 shows the top view of this filter with $d_1 = d_3 = 16.53$ mm, $d_2 = 17.72$ mm and the width of diaphragms $g_1 = g_4 = 16.53$ mm and $g_2 = g_3 = 7.50$ mm. We compact this filter using $N = 11$ alternating dielectric and ferrite layers for all three hollow sections d_1, d_2 and d_3 considering compactness ratio $K = 5/4, 4/3, 3/2$ and 2. Fig. 13 compares the amplitude of S_{21} of both air-filled waveguide and compacted DFW filters, obtained from the full-wave analysis utilizing finite element method of the HFSS software. It is seen that the frequency responses of the compacted filters are identical to that of the air-filled filter unless they have been stretched. This stretching is due to the difference between the slop of phases of DFWs and hollow waveguides with respect to frequency as in (13), and it is decreased as the compactness ratio K is decreased, or the desired frequency (with respect to the cutoff frequency) is increased. Because of the stretching effect of DFWs, we have to first design the waveguide

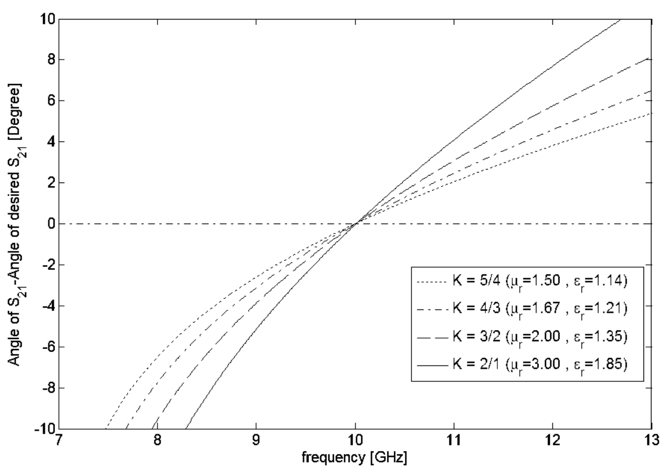

Figure 11. The difference between the phase of S_{21} of DFW and that of hollow waveguide for $k_h d_h = 90°$, $f_0 = 10$ GHz and $N = 11$.

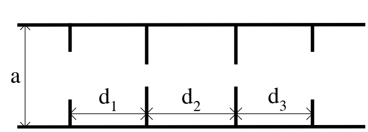

Figure 12. The top view of an air-filled waveguide bandpass filter with inductive diaphragms.

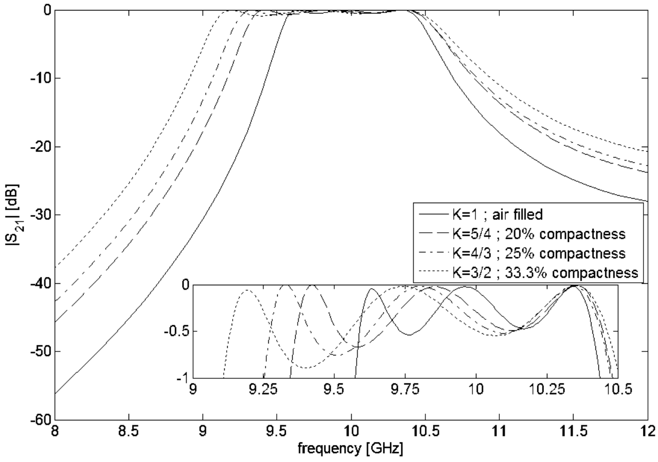

Figure 13. The simulation results of both air-filled waveguide and compacted DFW bandpass filters considering $N = 11$.

devices with narrower bandwidth than the desired bandwidth and then compact them.

Using the above examples and results, one may be satisfied with the performance of the proposed method to compact waveguide devices.

4. CONCLUSION

A method was proposed to compact narrowband previously designed waveguide devices. Some relations were obtained for the permittivity and permeability of dielectric and ferrite layers. The usefulness of the proposed method was verified using some theoretical and simulation examples. It was observed that the introduced compacting method could be used for narrowband waveguide devices. The performance of the resulted compact structure at desired frequency becomes better if the electrical length of hollow waveguide decreases; the number of dielectric and ferrite layers is increased; the desired frequency decreases, or the compactness ratio is decreased. Of course, the frequency responses of the compacted filters are identical to that of the air-filled filter unless they have been stretched. This stretching is decreased as the compactness ratio K is decreased, or the desired frequency (with respect to the cutoff frequency) is increased. Because of the stretching effect of DFWs, we have to first design the waveguide devices with narrower bandwidth than the desired bandwidth and then get to compact them.

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