Design of Planar Microwave Bandstop Filters by Control of Wave Propagation on Nonuniform Transmission Line

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Abstract

A design method is newly presented for the creation of planar type filters which shapes will be a nonuniform transmission line with continuously varying characteristic impedances. Those filters yield frequency characteristics of the bandstop filter in transmission, and have not any discontinuities. The design is achieved by the control of wave propagation in view to both reflection and transmission properties on the dispersive nonuniform line within the desired frequency band. The developed algorithm is based on optimized perturbations of inherent nulls of lobes from the solution of first-order nonlinear differential equation for reflection coefficients of the nonuniform line, and an appropriate distribution of reflection coefficient which affects to filter length and bandwidth. Practical measurements for designed wideband microstrip bandstop filters show very good agreement with theoretical results.

I. Introduction

Filters at microwave and millimeter wave frequencies require accurate theoretical design and precision fabrication. It is necessary to avoid the difficult task of physical fine tuning at those frquencies. To meet these criteria, planar or quasi-planar circuits have been widely studied in recent years. The conventional filter theory relies on synthesis to determine what properties for S_{21} must have to represent a network composed of L's and C's, whereby the nature of the transfer function is specified and the corresponding network is determined as a reactance network including some discontinuity[1]. However, little attention has been paid to the full potential of planar nonuniform filters which have not any discontinuities. The microstrip bandpass filter in reflection with a continuously varying width has been constructed using the inverse scattering method which are achieved by solving the Gel'fand–Levitan–Marchenko integral equation for a reflection coefficient of nonuniform transmission line[2]. However the design procedure is limited to the static TEM case and can't provide the complete prediction of transmission behavior at out-of-band because the result has unpredictable out-of-band ripple caused by truncation of the filter's impulse response.

In this paper, a new design method for realizing the planar filter is presented providing a generalized theory for the construction of continuously varying nonuniform line in dispersive media. This is carried out by an optimized perturbation of inherent nulls of lobe pattern representing the input reflection coefficient when a load is attached to the nonuniform transmission line. First of all, we introduced and modified Taylor's synthesis[3] of line source aperture distributions that yield antenna radiation patterns with Dolph-Chebyshev lobes within the specified region. The Taylor pattern is compatible with a lobe shape which is one of the various lobe patterns on the nonuniform transmission line, and representable as a canonical product of factors whose roots are nulls of the lobe pattern. The problem of designing a line source to yield a pattern with a prescribed sidelobe level envelope has been studied by Hyneman[4], who developed a perturbation procedure based on the log derivative of the envelope function. The concept led to computable shifts in the pattern nulls which caused an approximation to the requisite pattern. Accordingly, it is possible to control over the lobe heights representing magnitudes of reflection coefficients. The present paper also pertains to the general problem of arbitrary control of the individual lobe heights, and uses a Taylor pattern as the starting point. The control is achieved by the optimization technique developed for the appropriate null shift which affects to prescribed lobe heights including the full reflection. The technique is fundamentally different from that used by Hyneman. The characteristic impedance profile is found from inter-relations between Fourier transform pair[5] supporting the non-TEM mode and the optimized lobe pattern. The Fourier transform pair is derived from the series solution of first-order nonlinear differential equation of nonuniform line, and utilized as a tool for the lobe control and the determination of the filter profile. The all program codes for simulation

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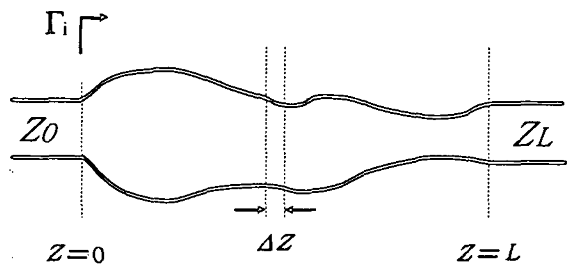


Fig. 1. Schematic representation of arbitrary nonuniform transmission line.

and design were developed by ours. Finally the wideband microstrip bandstop filters was constructed. The method was validated by the experiment.

Ⅱ. Filter Design Algorithms

Let the uniform guide with constant characteristic impedance Z_0 be connected to another uniform guide with characteristic impedance Z_L through the nonuniform line with a gradual transition which operates as a filter to reflect a specified frequency band (Fig. 1). The problem of interest is to find continuously varying characteristic impedances of the line satisfying the prescribed input reflection coefficients at each frequencies. From a forward traveling wave and a backward traveling wave with phase constant $\beta(\omega, z)$, the following first-order nonlinear differential equation for the reflection coefficient on the lossless dispersive line is obtained[5].

$$\frac{d\Gamma(\omega,z)}{dz} - 2j\beta(\omega,z)\cdot\Gamma(\omega,z) + \frac{1}{2}(1-\Gamma^2(\omega,z))\frac{d\ln\overline{Z}(\omega,z)}{dz} = 0 \quad (1)$$

In this expression, $\overline{Z}=Z/Z_0$ is the normalized characteristic impedance of the line which is frequency- and distance- dependent. We now consider the solution of (1) in two cases; in the first Γ is large value including the full reflection and in the second case Γ is small value.

Case 1: large values for Γ

The equation (1) does not have a closed form solution. Berquist [6] has reported a series solution for (1) valid at arbitrary load conditions. Under matched conditions, the solution is given by the expression, in the context of this paper,

$$\Gamma = \exp[2j\int_0^z \beta(\omega,z)\,dz\,]\cdot\frac{K_1+K_3+K_5+\cdots}{1+K_2+K_4+\cdots} \quad (2)$$

where

$$K_1 = \int_0^z f_1(\omega,z)\,dz$$
$$K_2 = \int_0^z f_2(\omega,z)\cdot K_1\cdot dz$$
$$K_3 = \int_0^z f_1(\omega,z)\cdot K_2\cdot dz$$
$$K_4 = \int_0^z f_2(\omega,z)\cdot K_3\cdot dz$$
$$\vdots$$
$$f_1(\omega,z) = \frac{1}{2}\frac{d\ln\overline{Z}(\omega,z)}{dz}\cdot\exp[-2j\int_0^z\beta(\omega,z)\,dz]$$
$$f_2(\omega,z) = \frac{1}{2}\frac{d\ln\overline{Z}(\omega,z)}{dz}\cdot\exp[\,2j\int_0^z\beta(\omega,z)\,dz]$$

It is assumed that only hybrid fundamental mode of propagation[7] exists on the line. The solution should also be applied to a types of transmission lines which vary so smoothly that the field structure along the line is not perturbed to such an extent that some discontinuities exist. That problem is solved by an proper distribution of reflection coefficient as explained later.

Case 2 : Smaller values for Γ ; $\Gamma^2 \ll 1$

Under this assumption, (1) reduces to

$$\frac{d\rho}{dz} - 2j\beta\rho + \frac{1}{2}\frac{d\ln\overline{Z}}{dz} = 0 \quad (3)$$

which can be solved for ρ_i

$$\rho_i(\omega,L) = \exp[2j\int_0^L\beta(\omega,z)\,dz]\cdot\int_0^L\frac{1}{2}\frac{d\ln\overline{Z}(\omega,z)}{dz}\cdot\exp[-2j\int_0^z\beta(\omega,\zeta)d\zeta]\,dz \quad (4)$$

Here ρ represents Γ in the case of $\Gamma^2\ll 1$, and ρ_i is the input reflection coefficient. The magnitude of the equation (4) corresponds to the term K_1 in (2). That has inherent lobes when the nonuniform line is used for transition section between the input and output. The examples are the exponential or Chebyshev tapered lines etc. for matching. In other words, the conventional way of solving problems on tapered transmission lines only for the impedance matching is to use an equation that corresponds only to the term K_1. If the larger reflection for filters is required within the specified frequency band, it is possible to search the complete solution from (2) using K_1. Here the null points of lobes of (2) is not changed from those of K_1 because magnitudes of f_1 and f_2 are same and phases have the reverse relationship each other. Accordingly, $|\rho_i|$ can be utilized for calculation of nonlinear differential equation of nonuniform line for the large reflection. If the input reflection coefficient is small($\Gamma_i^2 \ll 1$), the so-called Fourier transform pair[5] which is derived by equation (4) is well applied for determination of impedance profile along the transmission line. Upon making the substitution $p=2\pi(z/L-1/2)$ for z in (4), we obtain the following Fourier transform pair in terms of frequency- and distance-dependent characteristic impedances along the dispersive nonuniform line.

$$f(u) = \int_{-\pi}^{\pi} g(p)\,e^{-jpu}\,dp \quad (5)$$

$$g(p) = \frac{1}{2\pi}\int_{-\infty}^{\infty} f(u)\,e^{jpu}\,du \quad (6)$$

$$= \frac{1}{2} \frac{d \ln \overline{Z}(\omega, z)}{dp} \tag{7}$$

where

$$u = \frac{1}{\pi} \int_0^L \beta(\omega, z)\, dz \tag{8}$$

Here $f(u)$ is the magnitude of (4) in the u domain and u is related to the electrical length on the dispersive line. The impedance profile is extracted from (6) and (7) for the impedance matching section in the conventional way using the $f(u)$ specified as exponential or Chebyshev frequency responses.

However, since the Fourier transform pair can not be utilized directly in the case of large reflection coefficients ($\Gamma_i \rightarrow 1$), we make the pair only a tool for lobe control. Upon considering this concept, the controlled lobe heights of the $f(u)$ may be larger than 1. The null points and magnitudes of the series solution in the case of the large reflection consist with those calculated by (4) in the frequency domain and (5) in the u domain. But the lobe height calculated by (1) is suppressed under 1 even if the controlled lobe height using (4) and (5) is higher than 1. The inherent lobe schematic of (5) is analogous to the Taylor's procedure of the continuous line source for antenna pattern synthesis [3, 4] in view to the Dolph - Chebyshev pattern converted from the basic sum pattern $\sin \pi u / \pi u$, which results from the uniform excitation of the line source where $u = (2a/\lambda)\cos\theta$, a is the half-length of the source, and θ is the pointing direction from endfire. First of all, we should like to introduce the control of nulls of lobes because shifts of null points affect to lobe heights to be used for frequency characteristics of the filter. If the frequency response having arbitrary lobe height is required, it is difficult to calculate (5) and (6) because $Z(w,z)$ representing the nonuniform profile is still not known. Accordingly for the lobe height control, we introduce the line-source method modified for lobes of transmission lines instead of antenna patterns. The distribution function $g(p)$ may be zero outside the range $|p| > \pi$. This restriction corresponds to the physical requirement that $d \ln \overline{Z}/dz$ be different from zero only in the range $0 \le z \le L$. So, in the interval $-\pi \le p \le \pi$, $g(p)$ can be expanded in a Fourier series $g(p) = \sum_{n=0}^{N} (A_n \sin np + B_n \cos np)$ with the assumption that all real coefficients for $n > N$ are zero[5]. In this case we see from (5) that $f(u)$ must be in the form $f(u) = f_r(u) + jf_i(u)$ with $f(0) = 2\pi B_0$ and $f(n) = \pi(B_n + jA_n)$, $n = 1, 2, \cdot\cdot\cdot, N$. This result is a statement of the well-known sampling theorem and states that $f(u)$ is uniquely reconstructed from a knowledge of the sample values of $f(u)$ at $u=n$. From viewpoints above mentioned, we generalized Taylor's procedure compatible with equation(4), and extracted null points by the optimization for fitting to the prescribed lobe height (If ripple responses are required, dips are also considered.). The procedure is carried out from the following equation.

$$f(u) = C \frac{\sin k\pi u}{k\pi u} \cdot \frac{\prod_{n=n'}^{N_s} \left(1 - \frac{u}{u_n - jv_m}\right)\left(1 + \frac{u}{u_n + jv_m}\right)}{\prod_{n=n'}^{N_s} \left(1 - \left(\frac{u}{kn}\right)^2\right)} \tag{9}$$

where $n \supseteq m$

$n = (n_1, N_1), (n_2, N_2), \cdot\cdot\cdot, (n_s, N_s)$

k ; scaling factor

(n_s, N_s) is the subgroup to be optimized with the farther out lobes decaying in height according to the coefficient C which determines the lobe shape in the usual case of $u_n = 1, 2, 3, \cdot\cdot\cdot$, $v_n = 0$. u_n is the null point to be extracted and v_n affects to dip position which makes ripples. $k = 1, 1/2, 1/4, \cdot\cdot\cdot$ don't affects to the $g(p)$ because the sampled values $f(kn)$ at kn is same as those of $f(n)$ at n, but may affects to the convergence of the optimization since the search range for perturbed null points is related. The equation (9) provides various kinds of responses in accordance with values of u_n and v_n. N is the number of the nulls to be optimized in the passband. Let $f_p(u)$ be p'th peak value and $f_d(u)$ be d'th dip value, and the error function by the logarithmic difference is defined by the least squre method as follows.

$$E_l(X) = \sum_{p,d=n_s}^{N_s} \left(|\ln(f_p(X_l)/S_p(u))|^2 + |\ln(f_d(X_l)/S_d(u))|^2 \right) \tag{10}$$

where

$$X = (U, V)$$

l is the iteration number, $p = n_s, \cdot\cdot, N_s$, $p \supseteq d$, $U = [u_{ns}, \cdot\cdot, u_{Ns}]$ and $V = [v_{ns}, \cdot\cdot, v_{Ns}]$. Minimization of E is achieved by updating U and V to reduce the difference between the performance $f_p(u)$ and $f_d(u)$ achieved at any u_n, v_n and the specifications $S_d(u)$, $S_d(u)$ representing the prescribed objective p'th peak value and d'th dip value, respectively. The iteration scheme for the minimization with the stopping condition $E < \varepsilon$ is given by

$$X_{l+1} = X_l - a_l \cdot \mathbf{H}^{-1}(X_l) \cdot \nabla[E(X_l)] \tag{11}$$

where $0 < a_l < 1$ and **H** is the Hessian matrix[8]. The difficulty of inverse Hessian matrix at each stage of iteration in (11) can be easily overcome by Davidon -Fletcher -Powell algorithm. For all practical applications which have been investigated, excellent convergence to the desired pattern has been obtained.

Ⅲ. Numerical Results and Discussion

Fig. 2 shows the generalized strategy for the control of lobe heights in the u domain by perturbation of null points in the case of $Z_L/Z_0 = 3.57$ using (9). Here $n_1 = 4$, $N_1 = 10$ for null points, and $n_1 = 8$ for one dip in the case of $s = 1$. $S_p(u) = 0.05$, 0.05, 0.05, 0.1, 0.1, 0.05, 0.05 and $S_d(u) = 0.095$. The convergence

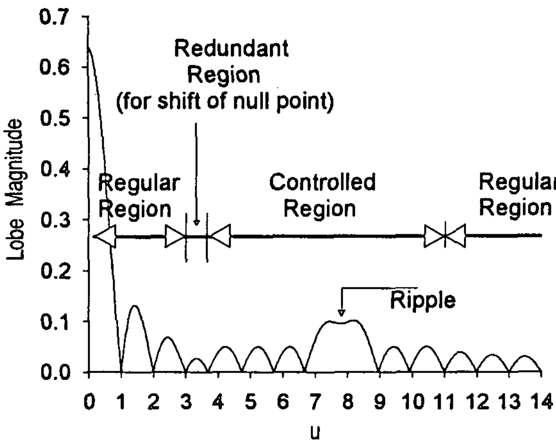


Fig. 2. Generalized strategy for lobe height control.

criterion adopted is $\varepsilon = 10^{-7}$. The results of optimized null points $u_{4\text{-}10}$ are [3.6879, 4.7332, 5.7295, 6.6717, 7.8004, 8.9589, 9.9236] and $v_8 = 0.4529$ for prescribed specifications. The figure shows that the approximate solution (5) of (2) wrongly predicts a reflection of 0.6363 at $u \rightarrow 0$ whereas the correct value is 0.5624, and that (5) is accurate as long as $f(u)$ is small. Accordingly, (2) must be used for high reflection and (5)-(9) are intended to use only for a tool of lobe height control. In Fig. 2 the first a half and two lobes represent exponential taper responses ($C = 1/2 \ln (Z_L/Z_0)$), and the controlled lobes involves the Dolph-Chebyshev characteristic with tolerable reflection coefficient 0.05 and 0.1 with a ripple. The named regular region may be inserted for reducing null points controlled, so the number to be optimized is reduced. But that strategy don't play a important role for the actual application for design of microwave filters. The redundant region must be inserted for providing the null shift achieved in the optimization process for the prescribed lobe heights. It is not necessary if $s = 1$ and $n_1 = 1$.

Fig. 3 (a) is an example of optimized high lobes fitted to prescribed responses (one high lobe, periodical lobe and lobe having ripples) in the u domain in the case of $Z_L/Z_0 \rightarrow 1$ ($Z_0 = 50\Omega$). For instance, the specification of periodical lobe is [10^{-4}, 10^{-4}, 0.05, 1.2, 0.08, 0.08, 1.2, 0.05, 10^{-4}, 10^{-4}]. Optimized null points are [0.3268, 0.6388, 0.8592, 2.3417, 5.1590, 5.7969, 6.4368, 9.2948, 10.7934, 11.2580]. Since the locus of reflection coefficients in the u domain cannot be distinguished from any phase constant, $u = \beta L/\pi$ may be used instead of (8)[9]. Therefore, although $\beta(w,z)$ has not been known before the impedance profile is synthesized, the u domain can be established by making values of βL varied within the involved range. Since those responses are equivalent to the equation (5), (6) can be used for calculating the impedance profile. Fig. 3 (b) shows the corresponding impedance profiles calculated by using the reflection coefficient distribution function $g(p)$. Fig. 3 (c) shows the corresponding frequency responses calculated by (2) when the microstrip is employed. Here, with the relations of (8) and $\beta(\omega, z) = \omega\sqrt{\varepsilon_{r,eff}(W/h, \varepsilon_r, \omega, z)}/c$ in which $\varepsilon_{r,eff}$ is the effective permittivity having dispersive characteristic and c is the light velocity, the frequency response is determined from the iterative procedure by considering the line length(L), bandwidth (B), and X which affects to successive lobe heights determining the reflection and transmission behaviors. Let S_m^p be the prescribed $m'th$ peak value

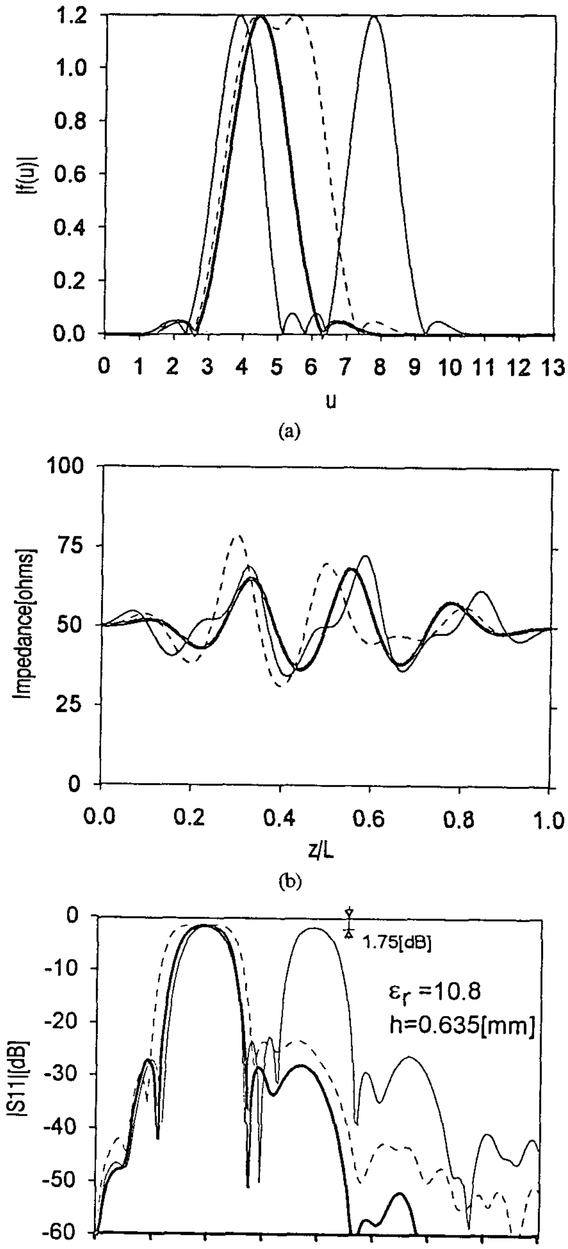


Fig. 3. (a) Examples of controlled lobes, (b) Corresponding impedance profiles, (c) Corresponding frequency responses (— : L = 25.1[mm], — : 20.8[mm], ------ : 28.3[mm]).

in the frequency domain, F^p_m the computed *m'th* peak value from (2), S_B the prescribed bandwidth and B the computed bandwidth at each iteration. The numerical representation for error function over the entire frequency range is then

$$E'(L,X) = [S_B - B(\beta, L, X)]^2 + \sum_{m=1}^{N} W_m \cdot [S^p_m(\omega) - F^p_m(\beta, L, f(X))]^2 \quad (12)$$

Here N is the involved lobe number within the specified range. The problem at hand is that of finding the values of L by updating the X for the minimum of E' to any degree of accuracy specified. W_m is the weighting function. That is mainly emphasized to high reflection lobe height, since the small lobe is almost consist with the result calculated by $f(X)$. In Fig. 3 (c), 3-dB bandwidths of one lobe and periodical lobe at the center frequency 10[GHz] are 30% and that of lobe with ripple is 50%. The lengths of microstrip line are also shown.

From the above viewpoint, the actual bandstop filter at 8[GHz] can be designed as shown in Fig. 4. The dotted line (60% bandwidth) has 11 lobe optimized for lobe heights (0.1, 0.1, 3.3, 0.1, 0.1, · · , 0.1) in the u domain. Those prescribed responses which are determined by the lobe control algorithm is converted to frequency domain by a series solution of nonlinear equation. Here the location of complete reflection in the frequency domain is determined by the location on the u domain, frequency dependent phase constants and line length. If the substrate is chosen, the line length is determined by taking account of the bandwidth and center frequency. The solid line (30% bandwidth) have 18 lobe heights (0.1, 0.1, 0.1, 0.1, 0.1, 3.3, 0.1, · · , 0.1) optimized. From the simulated result shown in Fig. 4, low lobes affect to the location of the high lobe and the distribution of reflection coefficient which directly affects to the impedance profile. The line length of dotted line is shorter than that of solid line. But the realization of the filter is difficult since the impedance profile have the possibility of generating some higher order mode and the strip width is very narrow. The reason is that the distribution of reflection coefficient is small.

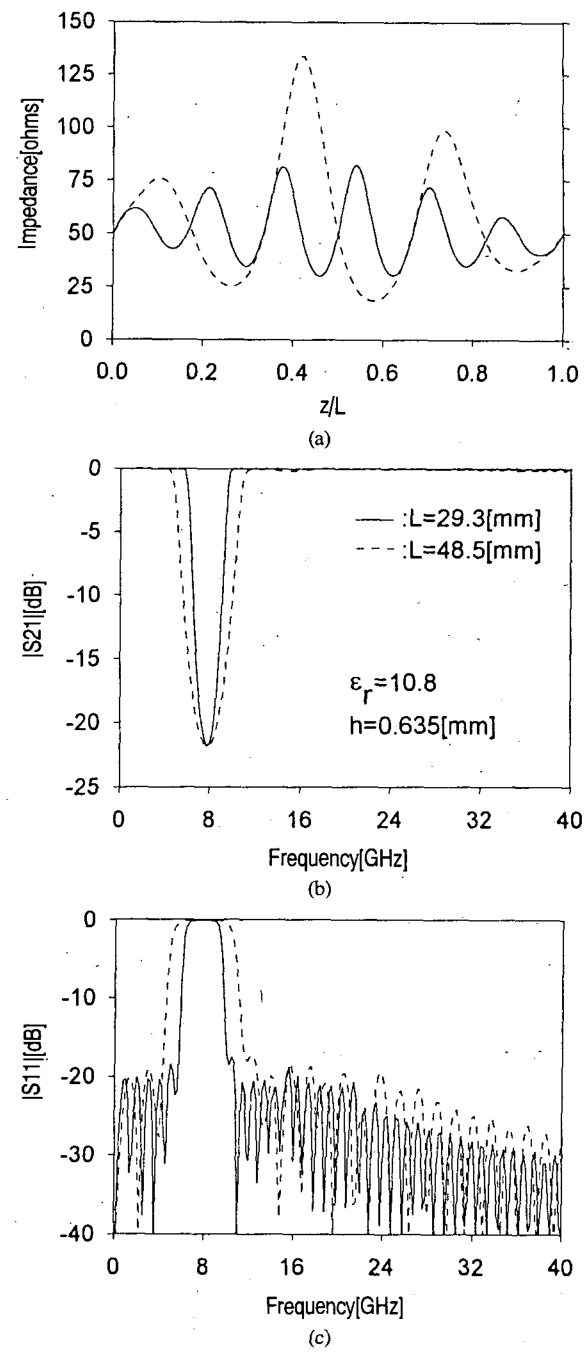
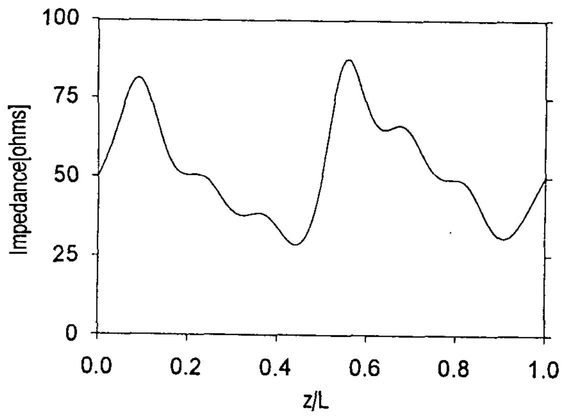


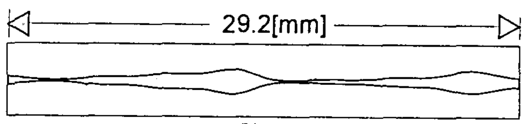
Fig. 4. Calculated impedance profiles and frequency responses for 8 [GHz] bandstop filters.

IV. Experimental Results

The periodical bandpass filter in reflection for the selective reflection amount -1.75[dB] and B = 1.5[GHz] was designed as shown in Fig. 5. The controlled lobes in the u domain is 11 (0.08, 1.2, 0.08, 1.2, 0.08, 1.2, 0.08, 0.05, · · , 0.05). Optimized null points are [0.0448, 0.8690, 2.9433, 3.3103, 5.2214, 5.5948, 7.7498, 8.4598, 9.2370, 10.1274, 11.0591]. That was fabricated in RT/ Duroid 6010 (ε_r = 10.8, h = 0.635 [mm]) and measured. The conductor pattern was etched onto the board using standard PCB fabrication techniques. The bandstop filter with 50% bandwidth at center frequency 10 [GHz] was also designed, fabricated and measured (Fig. 6). The controlled lobes are 15 (0.1, 0.1, 0.1, 3.3, 0.1, · · , 0.1). Optimized null points are [0.0513, 1.2725, 2.1214, 2.7124, 5.5911, 6.1809, 7.0180, 7.9430, 8.9054, 9.8874, 10.8816, 11.8840, 12.8935, 13.9100, 14.9371]. The conductor pattern is shown in Fig. 6(b). The computed results show that the variation of impedances along the filter is smooth and the strip width is in the permissible range in view to physical realization and hybrid fundamental mode propagation. The agreement between the theoretical and measured response is



(a)



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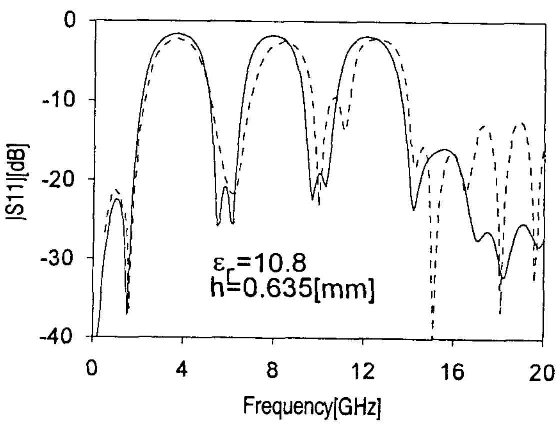


(c)

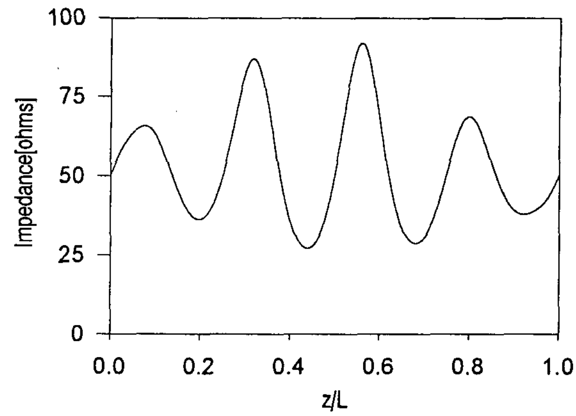
Fig. 5. (a) Impedance profile, (b) Conductor pattern, (c) Frequency responses of periodical bandpass filter in reflection (— : calculated, ····· : measured).

excellent except the error shown in the right corner in Fig. 6(d). The inconsistency is due to the error of Bolinder's formula[10]. The results was also validated from two port analyses by modelling the designed filter to equi-length 500 segments.

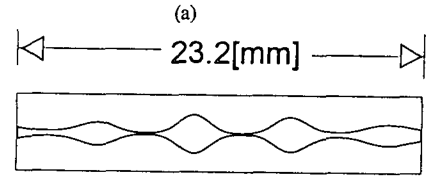
A similar design procedure may be followed for the synthesis of millimeter wave filters since very accurate results with a wide range of applicability for proper compensation for dispersion have been reported[11, 12]. And if the substrate with high dielectric constant is used, the filter length can be remarkably shorten since the phase constant affects to the length as shown in (8).

V. Conclusions

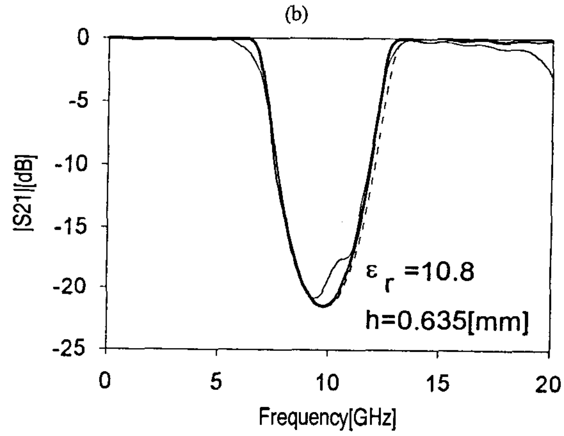
A new design method for microstrip bandstop filters has been presented, aimed at making it suitable for the construction of filters with continuously varying strip width. The method was based on the efficient lobe height control which is carried out by optimized perturbation of inherent nulls of lobes representing



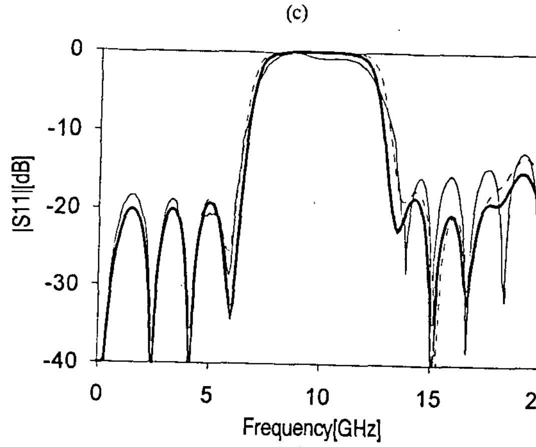
(a)



(b)



(c)



(d)

Fig. 6. Characteristics of designed 10[GHz], 50% bandwidth bandstop filter. (a) Impedance profile, (b) Conductor pattern, (c) and (d) frequency responses. (— : calculated, — : measured, ····· : modelled by 500 equi-length segments).

input reflection coefficients including the full reflection. Associated with an optimization procedure, this approach provides the generalized theory for the synthesis of the planar nonuniform line for prescribed reflection in dispersive media. Filters designed by the present method have advantages over other methods : 1) feasibility of continuously varying impedance profile, 2) simple control of wave propagation introducing the impedance analysis, 3) a faithful frequency response over a wide band. Periodical bandpass filter in reflection for generality and 10[GHz] bandstop filter were designed, fabricated and tested. The measured results are in good agreement with the results obtained by the proposed method.

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