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공간적응형 소스 분포 합성법을 사용한 임의의 반사응답을 갖는 전송선로 설계

(Design of Transmission Lines with Arbitrary Reflection Responses Using Synthesis Method for Spatially Adaptive Source Distribution)

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요 약

임의 복사패턴을 만족시키는 배열안테나 소스 전류분포 합성시 사용하는 Woodward-Lawson 샘플링법은 우함수형 복사패턴 합성에 주로 사용되어 왔다. 본 연구에서는 이를 확장하여 기함수형 패턴도 만족시키는 비선형 소스 분포함수의 최적합성법을 제시하고, 이를 임의의 반사특성을 갖는 단일 및 결합선로의 불균일 모드 임피던스 프로필 합성에 동시에 적용하였다. 이 최적합성법은 주파수영역 반사패턴에 내재된 복소 null점의 최적 섭동에 기본을 두고 있다. 제어된 반사패턴의 표본값으로부터 분산특성을 갖는 임피던스 프로필이 직접 계산되므로서 기존의 불균일선로 합성법보다 매우 간단함을 보였다. 그리고 서로 다른 임피던스간의 정합을 위한 기존의 테이퍼선로 이론에 기초한 불균일선로 합성법들을 탈피하여, 서로 같은 임피던스간의 불균일선로 합성도 가능케하므로서 본 연구에서 제시한 방법은 일반성을 가진다. 임의 통과대역을 갖는 필터 설계에 적용, 분석하므로서 타당성을 보였다.

Abstract

In the synthesis of the current source distribution function of an array antenna with the arbitrary radiation pattern, the Woodward-Lawson sampling method has been mainly used for the synthesis of an even function lobe pattern. In this paper, the method is extended to the synthesis of the odd function pattern and then the optimum synthesis method for the nonlinear source distribution function is proposed. The proposed method is applied to the design of nonuniform transmission lines with arbitrary reflection responses. The both dispersive impedance profiles of single and coupled nonuniform lines with arbitrary reflection responses are directly synthesized by the sampled values of a reflected spectral pattern which is optimally shaped by a perturbation of its complex null positions, hence removing the conventional step-by-step segmentation process and global optimization routines. The control problem in the case that all of port impedances are identical is also solved. The generality of the proposed method is verified by a filter design with the controlled arbitrary passband

Key Wodrs : Transmission Line, Filter, Pattern Control

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I. Introduction

The study on nonuniform transmission lines (NTLs) in frequency domain has been mainly concerned with the tapered line theories which are based on the first-order nonlinear differential equation derived from the telegrapher's equation.[1~3] The design of dispersive impedance transformers and asymmetric couplers with the taper has been most commonly based on the step-by-step segmentation process.[4,5] However, the scheme is limited to a special case such as the exponential and Chebyshev taper. The symmetric tapered couplers have been typically designed by an optimization of the coupling coefficient which is parameterized for a special backward coupling.[6,7] In order to overcome the above limited, individualized and complex methods, this paper presents a generalized design scheme for dispersive single and coupled NTLs with arbitrary passband in reflection. In the transition problem between two impedances(Z_0 and Z_L), two methods are proposed: one is to determine a nonuniform transition for arbitrary reflected levels in the specified band, based on an adaptive control of a reflected wave: the other is to make the procedure satisfied in both cases that $Z_0=Z_L$ and $Z_0 \neq Z_L$ by using improved Woodward- Lawson sampling method in so-called line-source method.[8,9] The Woodward-Lawson sampling method has been restricted to the even function line-source distribution making the asymmetrical impedance profile. Here we improved the method by extending to the case of odd function line-source distribution making the symmetrical impedance profile. In the scheme, the reflection properties of those lines are adapted to the prescribed pattern in the frequency domain by a perturbation of inherent complex null points in its features, and then the corresponding impedance profiles are directly determined by the sampled values of the adapted pattern. The generalized scheme is well suited to the design of dispersive single- and coupled-lines with arbitrary passband, without the step-by-step segmentation process and overall optimization routines. Since the scheme is basically established in the electrical length domain, the problem of dispersion treatment is also removed.

II. Generalized Scheme

Let the characteristic impedance Z_0 at $z=0$ be connected to another impedance Z_t^L at $z=L$ through the lossless dispersive line with a gradual transition represented by the frequency-dependent impedance, $Z_s(z,f)$. And letting the four ports of a coupled-line coupler be terminated in Z_0, the relationship $Z_0=\sqrt{Z_{0e}(z,f)\cdot Z_{0o}(z,f)}$ for $z \in (0,L)$ is preserved throughout the design, in which $Z_{0e}(z)$ and $Z_{0o}(z)$ are the frequency-dependent even- and odd-mode impedances, respectively. Since the coupler can be simplified to the two-port analysis in terms of the even mode characteristics,[10] the magnitude of the input reflection coefficient from the theory of reflections on a non-TEM lossless tapered line can be, in general, expressed as $|H_t(f)| \approx \tanh(|h_t(f)|)$, in which $h_t(f)$ is expressed as follows.[7]

$$h_t(f)=\int_0^L \frac{1}{2}\frac{d}{dz}\ln\left(\frac{Z_t(z,f)}{Z_0}\right)\cdot \exp\left(-j2\int_0^z \beta_t(z',f)dz'\right)dz \quad (1)$$

where β_t is the phase constant of the transition. The general index s denotes as the index in the case of the single line. In the case of -length coupled-line coupler with the system impedance Z_0, the index can be replaced by $0e$ representing the even-mode. First, we specify a design frequency as $f=f_0$ and treat the L with a dilation factor, in the context of this paper. Defining that $p=2\pi(z/L-1/2)$ and $u=\beta_t(f)\cdot L/\pi$ instead of $u=\int_0^L \beta_t(z,f)dz/\pi$, we obtain the following relationship over the range $|p| \leq \pi$.

$$h_t(u)=\int_{-\pi}^{\pi} g(p)\cdot \exp(-jpu)dp \quad (2)$$

$$g(p)=(d\ln(Z_t(p,f_0)/Z_0)/dp)/2 \quad (3)$$

Expanding the $g(p)$ as $g(p)=\sum_{n=0}^{N}(a_n\cos(np)+b_n\sin(np))$ by the $h_t(u)$ Woodward's idea,[8] the can be rewritten by the restricted set of sampling functions as follows.

$$h_t(u) = \sum_{n=0}^{N} \pi a_n (Sa(\pi(u-n)) + Sa(\pi(u+n))) - j\sum_{n=1}^{N} \pi b_n (Sa(\pi(u-n)) - Sa(\pi(u+n))) \quad (4)$$

where $h_t(n) = \pi(a_n - jb_n)$ and $h_t(0) = 2\pi a_0$. The real and imaginary part are even and odd in u, respectively. When $Z_t^L \neq Z_0$ (or $h_t(0) \neq 0$) and $b_n = 0$, the schematic of $|h_t(u)|$ is typical of a lobelike taper response which has the first half a lobe followed by full lobes with null points in the positive u domain. The lobe schematic is analogous to that of the usual pattern factor of the continuous line source antennas.[9] And then $g(p)$ is corresponded to the nonlinear distribution function of the line source. The problem we now are considering is how to utilize the full advantage of the restricted $h_t(u)$ in cases of $b_n \neq 0$ or $Z_t^L/Z_0 = 1$. We introduced the Taylor line source pattern[9] of $h_t(u)$ for a lobe control, and then modified that in order to activate the case that $b_n \neq 0$, based on the Orchard's ripple-making theory.[8] The result is as follows.

$$h_t(u) = A \cdot Sa(\pi u) \cdot \prod_{n=1, n \geqq m}^{N} \left(\frac{n^2}{n^2 - u^2}\right)\left(1 - \frac{u}{u_n - jv_m}\right)\left(1 + \frac{u}{u_n + jv_m}\right) \quad (5)$$

whose real and imaginary part are even and odd, respectively. That is $h_t(u) = h_{t,e}(u) + jh_{t,o}(u)$, where $h_{t,o}(0) = 0$. Thus, the framework of eqn.(5) is consistent with that of eqn.(4), showing the general complex form equivalent to the set form of sampling functions. u_n is the null or dip position in u. v_m causes a dip in position which makes ripples. The and u_n are v_m optimally perturbed for the $|h_t(u)|$ which has the individually prescribed N lobe heights with the furthest lobes exponentially decaying in level according to the coefficient $A = 1/2\ln(Z_t^L/Z_0)$.

On the other hand, if $Z_t^L/Z_0 = 1$, then $h_t(u) = 0$. This is trivial because of no reflection. So, it is required that $h_t(0) = 0$ and $h_t(u)$ has no any deviation against the regular pattern at $u \neq 0$. Thus, $h_{t,e}(u)$ must become odd. Once the alteration is done, $h_{t,o}(u)$ must become even and then the two derivations must be interchanged in position for maintaining the framework which is consistent with the need to have $g(p)$ be real. Letting the altered be $h_t^A(u) = h_{t,o}^A(u) + jh_{t,e}^A(u)$ and defining $h_t^A(u) = jh_t^A(u)$ yield $a_t^A(u) = a_{t,e}^A(u) + jh_{t,o}^A(u)$. After some algebra on the evenodd alteration in eqn.(4), we derived that $h_{t,e}^A(u) = h_{t,o}(u) + D_1$ and $h_{t,o}^A(u) = h_{t,e}(u) - D_0 - D_2$ where $D_1 = 2\pi \cdot \sum_{n=1}^{N} b_n \cdot Sa(\pi(u-n))$, $D_0 = 2\pi a_0\, S_a(\pi u)$ and $D_2 = 2\pi \cdot \sum_{n=1}^{N} a_n \cdot Sa(\pi(u+n))$. Thus, the following relationship is obtained.

$$h_t(u) \leftrightarrow g(p) = \sum_{n=0}^{N} (a_n \cos(np) + b_n \sin(np)), \text{ if } Z_t^L / Z_0 \neq 1$$
$$h_t'(u) \leftrightarrow g(p) = \sum_{n=1}^{N} (b_n^A \cos(np) + a_n^A \sin(np)), \text{ if } Z_t^L / Z_0 = 1 \quad (6)$$

where b_n^A and a_n^A are the updated $b_n(= -h_{t,o}(n)/\pi)$ and $a_n(= h_{t,e}(n)/\pi)$ of $h_t(u)$ which contributes to a prescribed $|h_t^A(u)|$. Letting the general notation for $h_t(u)$ and $h^A(u)$ be $h_t^G(u)$ and then letting $h_{t,p}^G(u)$ be th peak value and $h_{t,d}^G(u)$ be the th dip value, the error function is defined by the least square method as follows.

$$E(\mathbf{X}) = \sum_{\substack{p,d=1 \\ p \geqq d}}^{N} \left(\left|\ln(h_{t,p}^G(\mathbf{X})/S_p)\right|^2 + \left|\ln(h_{t,d}^G(\mathbf{X})/S_d)\right|^2\right) \quad (7)$$

where $\mathbf{X} = (\mathbf{U}, \mathbf{V})$ in which $\mathbf{U} = [u_1, \cdots, u_N]$ and $\mathbf{V} = [v_1, \cdots, v_N]$. Minimization of E is achieved by updating $\mathbf{X}$ to reduce the logarithmic difference between the performances $(h_{t,p}^G, h_{t,d}^G)$ during the updating process and the specifications (S_p, S_d) which represent the prescribed objective pth peak value and th dip value. The iteration for minimization with stopping condition $E < \varepsilon$ is along with the Davidon-Fletcher-Powell algorithm.

Ⅲ. Applications and verifications

For the generalized strategy, we consider that

$N=10$, $m=3,7,8$, $S_p=0.05$ for $p\in(1,N)$ and $S_d=0.04$ for $d\in m$. When $Z_t^L/Z_0=2$ for a single-line taper or tapered asymmetric coupler, the $|h_t(u)|$ adapted by the Davidon-Fletcher-Powell algorithm ($\varepsilon=10^{-6}$) and the synthesized profile $Z_t(z)$ by the distribution function $g(p)$ calculated from the sampled value $h_t(n)$ are shown in Fig.1. The peaks in the case of $N\succ 10$ are exponentially decayed as expected. Considering the tapered asymmetric coupler, Z_t^L is corresponded to Z_{0e}^L which is the even mode characteristic impedance at the point of the strongest coupling.[10] The complete Dolph-Chebyshev (0.05) taper profile is also shown in Fig.1 for comparison. Assuming $Z_t^L/Z_0=1(h_t^A(0)=)$, for the rapid convergence, the nulls and dips are optimally perturbed for $|h_t(u)|$ in the case that $|h_t^A(u)|$ and then the resulting complex nulls are used as the initial values for . The simulated results in cases that $Z_t^L/Z_0=1.001$ and $Z_t^L/Z_0=1$ are shown in Fig.2. The inset is the magnified figure at the small u. Fig.3 shows peaks and dips of the above controlled reflection coefficients in the complex plane. The optimally perturbed u_n and v_n are summarized in Table 1. And resulting coefficients in (6) are summarized in Table 2.

표 1. 미리 설정한 $|h(u)|$와 $|h_t^A(u)|$에의 적응을 위한 최적화된 u_n과 v_n

Table 1. Optimized u_n and v_n for the adaptation to prescribed $|h(u)|$ and $|h_t^A(u)|$.

	$Z_t^L/Z_0=2$	$Z_t^L/Z_0=1.001$	$Z_t^L/Z_0=1$
u_1	1.019277	0.059255	0.092440
u_2	1.809379	1.499813	1.290781
u_3	2.723735	2.532773	2.469814
u_4	3.683074	3.546933	3.280433
u_5	4.729072	4.627844	4.407801
u_6	5.788947	5.712918	5.524132
u_7	6.797376	6.739518	6.663975
u_8	7.739724	7.692069	7.637825
u_9	8.760562	8.724529	8.578154
u_{10}	9.857528	9.831852	9.743455
v_3	0.283660	0.308753	0.673465
v_7	0.317975	0.324920	0.316229
v_8	0.321324	0.327680	0.466881

From the above examples, it is assured that the arbitrary reflection coefficient can be easily controlled and the corresponding impedance profile be directly

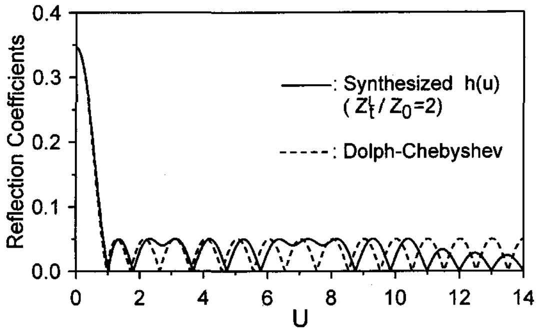
표 2. 분포함수 $g(p)$의 계산을 위한 계수들의 시뮬레이션 결과

Table 2. Simulated results of coefficients for distribution functions $g(p)$.

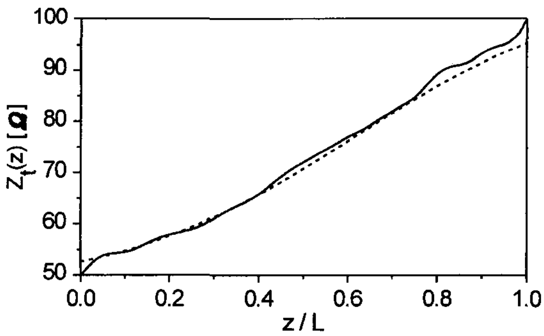
n	$Z_t^L/Z_0=2$		$Z_t^L/Z_0=1.001$		$Z_t^L/Z_0=1$	
	b_n $(\times 10^{-3})$	a_n $(\times 10^{-3})$	b_n $(\times 10^{-3})$	a_n $(\times 10^{-3})$	a_n^A $(\times 10^{-3})$	b_n^A $(\times 10^{-3})$
0	nil	55.1589	nil		nil	nil
1	-0.2049	1.8702	2.1222	-15.3455	1.2085	-4.3928
2	-3.3081	8.5875	-7.6118	13.5165	-10.3024	8.5499
3	-10.4403	-10.9848	-8.2946	-13.5911	-5.8911	-4.5050
4	1.5288	12.8287	1.6306	15.2839	3.7288	10.4589
5	0.9116	-11.4798	1.2247	-14.0434	-0.7160	-13.9997
6	-3.4521	8.2539	-4.7604	10.3311	-6.1508	14.7211
7	-9.3669	-10.3812	-8.2554	-12.0587	-6.0724	-14.4932
8	14.1668	4.3773	14.228	5.5642	15.1381	2.9064
9	-4.8284	-8.8093	-5.3655	-9.7577	-9.5877	-11.9952
10	2.0990	5.9675	2.4666	6.8844	5.3742	10.3714

constructed by the sampled values. Furthermore, even if the port impedances are identical ($Z_t^L/Z_0=1$), the control technique is well suited for any reflection. For a practical example, assuming that is (0.1, 0.1, 0.1, 0.1, 0.1, 0.883, 0.07, 0.07, 0.883) for $N=9$ and $v_n=0$, a periodic 3-dB ($H-t^A(u)$) bandpass filter in reflection can be designed. Since $v_n=0$, $b_n=b_n^A=$ in eqn.(6). The results are shown in Fig.4 (a) and (b). For all practical applications which have been investigated, excellent convergence to the desired lobe pattern has been obtained.

The problem at hand is that of finding the frequency responses. Since we have let $u=\beta_t(f)L/\pi$, the schematic of $|h_t^G(u)|$ may be contracted or dilated by L in the frequency domain, while the lobe peaks of $|h_t^G(f)|$ are accordance with those of $|h_t^G(u)|$.



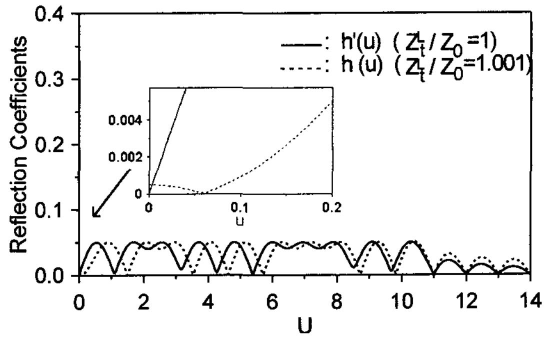
(a)



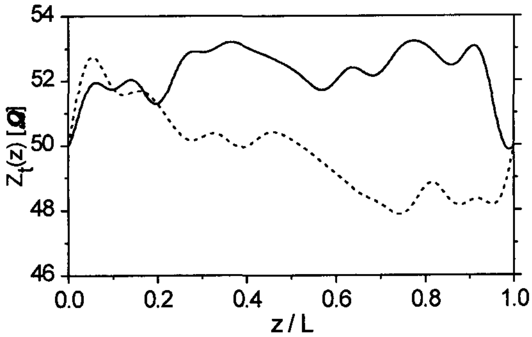
(b)

그림 1. 일반화된 Lobe 제어.
(a) 제어된 반사계수.
(b) 해당 임피던스 프로필

Fig. 1. Generalized stratedgy for lobe control.
(a) Controlled reflection coefficients.
(b) Corresponding impedance profiles.



(a)



(b)

그림 2. 동일한 두 임피던스간의 변환.
(a) 제어된 lobe
(b) 해당 임피던스 프로필.

Fig. 2. Transition determinations between two identical impedances.
(a) Controlled lobes.
(b) Corresponding impedance profiles.

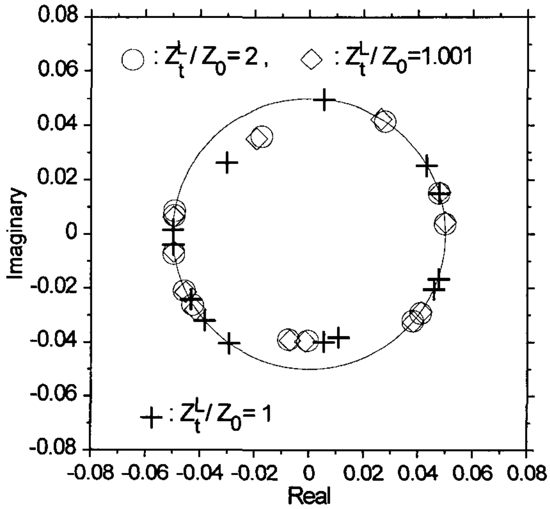


그림 3. 복소평면에서 제어된 lobe의 피크값과 딥값.

Fig. 3. Peaks and dips of controlled lobes in complex plane.

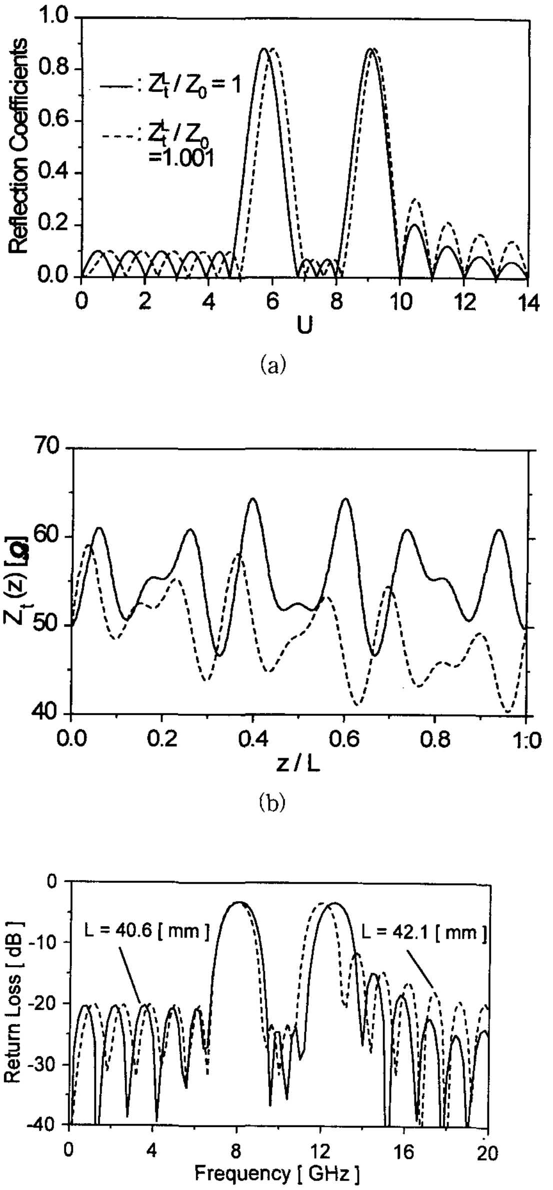


그림 4. 주기적인 반사형 대역통과예
(a) 제어된 반사계수.
(b) 합성된 해당 임피던스 프로필.
(c) $\varepsilon_r=10.2$, $h=0.635$ mm인 마이크로스트립상에 구현했을 경우의 계산된 주파수응답

Fig. 4. Examples for periodical bandpass in reflection.
(a) Controlled reflection
(b) Synthesized corresponding impedance profiles.
(c) Evaluated frequency responses of implemented microstrip lines $\varepsilon_r=10.2$, $h=0.635$ mm

Accordingly, the line length L as the dilation factor can be easily optimized by generating an error function of frequency performances such as the bandwidth and center frequency, and by slightly changing the inserted lobe number and peaks in the domain specification. Let's consider the just above example. The chosen peaks and inserted number of lobes acts as the centering in 8 GHz and 12 GHz in the case of $Z_t^L=Z_0=1.001$. Fig.4 (c) is the frequency response calculated by the usual two-port analysis performed on the cascaded equi-length segments which consist of the microstrip line ($\varepsilon_r=0.635$, $h=0.635mm$) satisfying the impedance profile shown in Fig.4 (b). Here, the closed-form design formula[11] has been used. The exact frequency centering in the case of $Z_t^L=Z_0=1$ may be possible in a similar manner.

Let's consider the coupled-line 3-dB bandpass filter with 2 GHz bandwidth at 10 GHz, aimed at making all ports match the system. The adopted S_p are (0.13, 0.1, 0.1, 0.1, 0.1, 0.883, 0.1, 0.1) for $N=8$. The first lobe height has been chosen for the condition $Z_{0e}(z) \succ Z_0$ which has to be satisfied for the physical realization of coupler.[7] The $Z_{0e}(z)$ synthesized from the adapted $|h_{0e}^A(u)|$ is shown in Fig.5 (a). $|H_{0e}^A(f)|$ means the coupling factor C in the symmetric coupler.[6,7] The spacing $S(z)$ between side edges of two strips and strip width at the design frequency $f_0=10GHz$ have been optimized by adopting Kirschning and Jansen's formulas[12] under the coupler condition $Z_0=\sqrt{Z_{0e}(z)Z_{0o}(z)}$, and then the wiggly technique for the equalization of phase velocities of even and odd mode has been applied by introducing wiggle depth $D(z)$.[7] The results are shown in Fig.5 (b). The implemented conductor pattern on the RT/Duroid 6010 microstrip substrate ($E_r=10.2$, $h=0.635mm$)is shown in Fig.5 (c). Since the configuration is the symmetrical continuous coupler, the phase nature has the familiar quadrature property with good group delay.[6] And the integrated configuration has more advantage than a waveguide

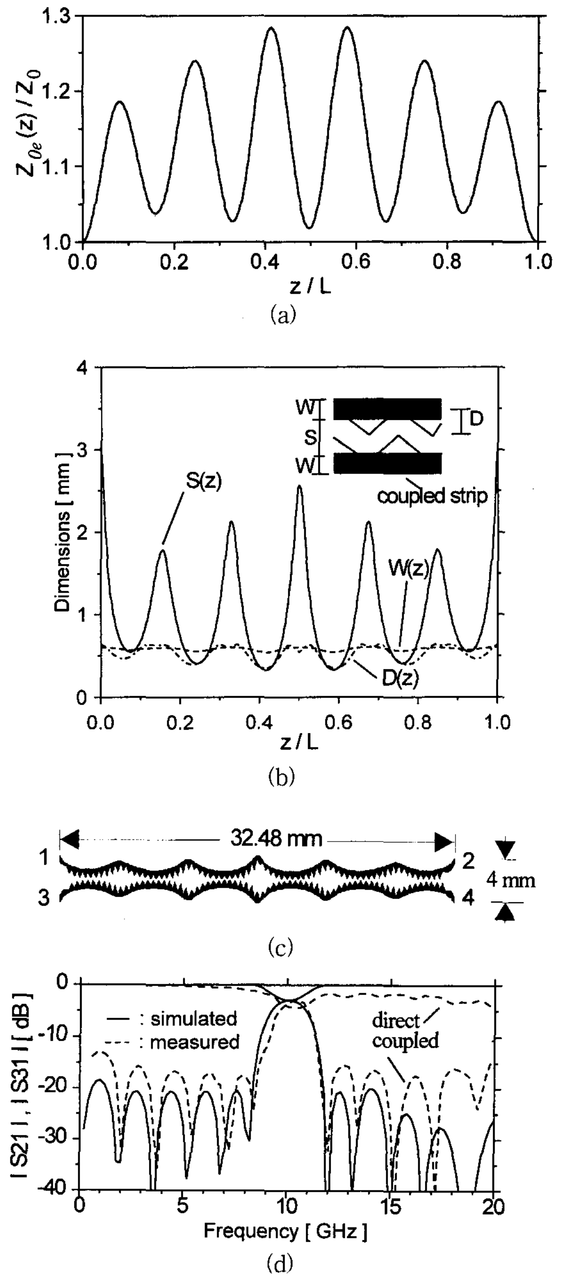


그림 5. 설계된 3dB 대역통과여파기와 성능. (a) 합성된 임피던스 프로필 (b) 마이크로스트립 기판 ($\varepsilon_r = 10.2$, $h = 0.635$ mm)상에 구현된 치수. (c) 도체 패턴. (d) 결합기해석을 통한 계산값과 측정값.

Fig. 5. Designed 3 dB bandpass filter and its performances. (a) Synthesized impedance. (b) Physical dimensions on microstrip. substrated ($\varepsilon_r = 10.2$, $h = 0.635$ mm). (c) Conductor pattern. (d) Calculated results by coupler analysis and measured results.

when the proposed technique is applied. The frequency characteristics have been calculated by the usual coupler analysis in which the S-parameter are converted from the chain matrix of the designed microstrip coupler with equi-length segmentation. The calculated frequency characteristics are compared with measured results in Fig.5 (d), showing good agreement.

Ⅳ. Conclusions

The single and coupled NTLs with arbitrary passband has been easily designed by a generalized scheme based on the optimum perturbation of inherent complex null points in a reflected wave, whether port impedances connected by a transition line are identical or not. The control algorithm has been developed from the shaped-beam pattern synthesis technique in the line source antenna design problem. The conventional taper theories focussed on the asymmetric transition between two different impedances has been improved by making the transition between same impedances be possible. Since the developed scheme is basically established in the electrical length domain, the problem of dispersion treatment can be automatically solved. Furthermore, since the dispersive impedance profiles can be directly obtained from the sampled values of controlled lobe pattern, the conventional step-by-step segmentation and repeated overall optimization process for a dispersive NTL can be removed.

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