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DESIGN OF MULTIBAND UWB FILTER BASED ON REFLECTED CHARACTERISTICS IN TIME DOMAIN

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Abstract—An automatic process to design a multiband filter in Non-uniform Transmission Line (NTL) form is presented. The proposed approach supports with Time Domain Reflectometry (TDR) technique instead of the traditional methods such as methods based on Inductor-Capacitor filters. The proposed method is described step-by-step and is illustrated by an example emphasizing key points. A critical analysis of this technique is done for emphasizing its limitations. For illustrating the design process, a multiband Ultra Wideband (UWB) filter rejecting two frequency bands is designed.

1. INTRODUCTION

Microwave filters are two-port networks used to control the frequency response at a certain point in a microwave system. They are present in any type of microwave communication, radar or test and measurement system. Many studies have been done for investigating various kinds of filters presenting typical responses including low-pass, high-pass, bandpass, and stopband characteristics and different forms [1].

With the emergent Ultra-Wideband technology, microwave filters gained again attention and a lot of effort has been put into developing planar microwave UWB filters operating over [3.1–10.6 GHz] and presenting high performances such as high selectivity, low losses, flat group delay and small circuit area. In particular, it is necessary to avoid all the interfere effects in the band of 2.4 GHz (used by almost public applications such as Wi-Fi, cell phone, wireless camera, webcam, microwave oven, etc.) and the band of 5 GHz (where operates

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WLANs (Wireless Local Area Networks), WIMAX (Worldwide Interoperability for Microwave Access), and ISM (Industrial-Scientific-Medical) systems). Furthermore, the world regulations are different and require different emission masks. For example, in contrast to FC's (Federal Communication Commission) single emission mask level over the entire UWB band, in Europe and Japan the radiation mask for indoor devices has two sub-bands: a low band ranging from 3.1 GHz to 4.8 GHz and a high band from 6 GHz to 8.5 GHz in Europ; from 3.4 GHz to 4.8 GHz and from 7.25 GHz to 10.25 GHz in Japan. In consequence, as one of the necessary components in a UWB communication system, UWB bandpass filters (BPFs) have attracted a lot of interest.

By rejecting several interfering bands, multiband UWB filters could take an important place in all UWB applications. In the literature, many topologies of UWB filters have been proposed [2]. Several topologies have been proposed including mainly filters based on multi-mode resonators with the assistance of coupling mechanism [3, 4], filters using cascaded circular or square ring resonators [5], microstrip filters implemented with cascaded short-circuited or open-circuited stubs [6], direct or indirect combination of a lowpass filter and a highpass filter [7, 8], and filters developed with hybrid transition of microstrip and coplanar waveguide structures [9]. If design guidelines for compact planar UWB filters based on computer simulations are often presented, few papers provide a synthesis theory. However some recent works give both design equations and guidelines providing a fast and easy implementation. In [10], a new topology of BPF based on interference techniques is introduced and both analytical equations and rules are given. The structure is formed by a short-ended coupled line coupler connected in parallel to a transmission line. However, an important step of tuning of the physical dimensions with a full wave simulator is required in order to reach the final topology. Recent studies, developed in Kanagawa and Saitoma Universities, proposed synthesis theories for UWB BPFs based on equivalent transmission line circuits [11]. Designs of UWB filters with different configurations including multi-mode step-impedance resonators combined with short-circuited and open stubs are described. Whatever the presented technique, the basic principle is always to specify the filter dimensions in order to satisfy the researched frequency gain response. In [12], a simplified time domain technique is presented for designing a UWB BPF for the band 3–10 GHz, built on a nonuniform transmission line (NTL).

In this context, this paper aims two objectives: to design UWB multiband filters, and for this, to propose an automatic process. The proposed filter design method is based on Time Domain Reflectometry (TDR) and Non-Uniform Line (NTL) technique. Unlike the traditional methods dealing with the Inductor-Capacitor (LC) components, the principle is to find out a resistance line by analyzing the reflected wave at the input of the transmission line to detect the characteristic of each discontinuity, which means to find out the various impedances of the filter line from its discontinuities. This approach had been appeared in some publications in 1990s [13–15] and studied also more recently [16], since this is a new potential method with many advantages (simple design, low cost, easy and reproductive fabrication, etc.). The proposed approaches have been improved by [17] that proposed a technique to reconstruct a nonuniform transmission line by using arbitrary incident waveforms. [18] completed this technique with the development of an algorithm simple and accurate enough for some practical applications. Although a low-pass filter was designed successfully as one of the first applications of this method [14], until now, there hasn't been yet an analytical method that allows calculating mathematically such kind of filter. Therefore in this paper, the method is completed by adding a totally automatic process from a specification of a desired filter to a NTL filter, then directly printed on electronic boards. In Section 2, the filter design process is presented; each step is explained and justified. An example of UWB multiband filter rejecting two bands of 2.4 GHz and 5 GHz. illustrates completely the method. In Section 3, the proposed method is evaluated from electric and electromagnetic simulations of designed filter characteristics. A critical analysis of the method is done in Section 4 for emphasizing its theoretical and physical limitations. Conclusion is given in Section 5.

2. FILTER DESIGN PROCESS

This part presents step-by-step the process to design a multiband filter. The different steps of this process can be illustrated as Figure 1. The first step is to determine the desired filter specification in term of the reflection coefficient S_{11}. This parameter is then converted to the corresponding reflected wave V_r. The following step is to analyze V_r to obtain the various impedances Z in the transmission line.

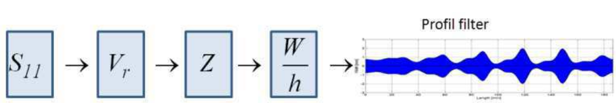

Figure 1. Filter design process.

Afterwards for given characteristic impedances and dielectric constant, the W/h ratio can be calculated. This ratio gives the dimensions of the microstrip line where W is the width of the conductor and h is the thickness of the grounded dielectric substrate. Finally, the result is a physical filter in NTL form which is considered as a various impedance chain.

2.1. Step 1: Determination of the Reflected Response V_r

First of all, a filter is represented by its transmission and reflection parameters, S_{21} and S_{11} respectively. The classic methods of microwave filters design such as the image parameter method and the insertion loss method are directly based on the exploitation of the transmission parameter specifications in frequency domain. However, the presented method focuses on the reflected wave, and so the reflection parameter is used to determine the filter. The ideal characteristics of the filter can be specified in terms of S_{21} parameter, and the reflected wave is then deduced. The coefficient S_{11} is determined in lossless condition as:

$$S_{11} + S_{21} = 1 \tag{1}$$

and the reflected wave $v_r(t)$ is easily obtained using one of the two following equations:

- in time domain (using the convolution operator "$*$"):

$$v_r(t) = s_{11}(t) * v_e(t) \tag{2}$$

- or in frequency domain:

$$V_r(f) = S_{11}(f) \cdot V_e(f) \tag{3}$$

  where v_e is the input wave in time domain. In the second case, an inverse Fourier transform allows the calculation of the reflected wave in time domain.

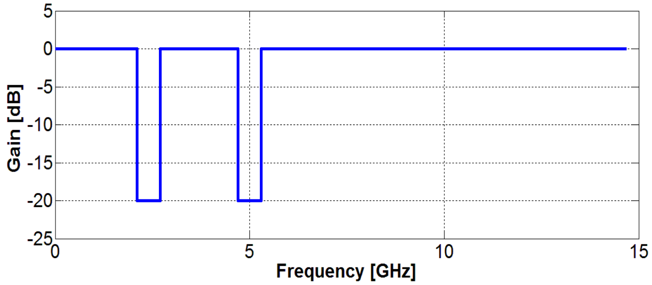

Figure 2. Transmission parameter S_{21}.

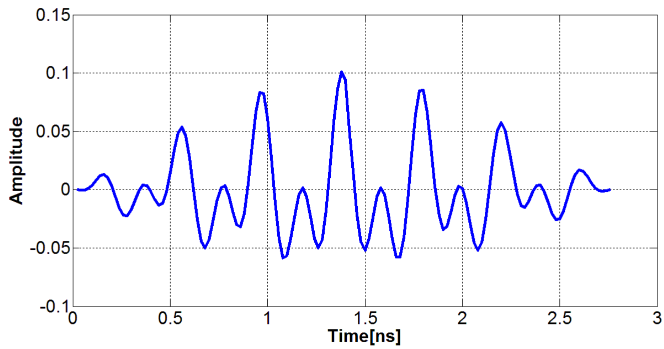

Figure 3. Reflected wave of the ideal filter.

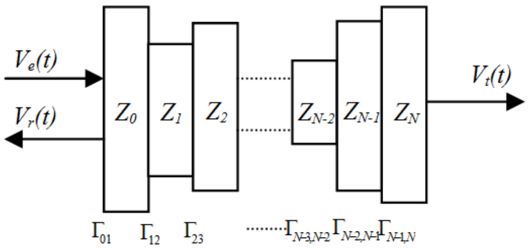

Figure 4. Example of a transmission line filter.

Figure 2 illustrates the S_{21} filter specification in which the bands of 2.4 GHz and 5 GHz are rejected. The corresponding reflected wave is represented in Figure 3.

2.2. Step 2: Determination of the Impedance Profile

The designed filter can be considered as a cascaded multi-section transmission line, i.e., a line of various impedances $Z_1, Z_2, \ldots, Z_{N-1}, Z_N$ connected by N discontinuities represented by the transmission and reflection parameters respectively $T_{i,\,i+1}$ and $\Gamma_{i,\,i+1}$ such as:

$$1 + \Gamma_{i,\,i+1} = T_{i,\,i+1} \tag{4}$$

Figure 4 illustrates the geometry of the NTL and the principle of impedance calculation. The injected wave v_e at the input end of the line is transmitted and reflected at each discontinuity; the wave coming back to the input end is the reflected wave v_r.

The analysis is based on the reflected wave at the input end after each $2n\tau$ seconds (with $n = 0, 1, 2, \ldots, N$ and τ the propagation time of wave along a section of line), the number N of sections, and the propagating waves. Assuming causality and considering all sections having same length l, τ is expressed as:

$$\tau = \frac{l\sqrt{\varepsilon}}{vT_e} \tag{5}$$

where v is the wave speed in the line, ε is the effective dielectric constant, T_e is the sample time. The length of the line is: $L = l \cdot N$. The parameters l and N play an important role in order to verify the initial specifications, and in the same time determine the size of the filter: the choice is a compromise.

The reflected wave can be expressed as follow:

$$V_r(2n\tau) = V_{r_{wf}}(2n\tau) + V_{r_{nwf}}(2n\tau) \tag{6}$$

$V_{r_{wf}}(2n\tau)$ is the wavefront which is transmitted without reflection to the farthest reachable discontinuity (the boundary between Z_n and Z_{n+1}) and $V_{r_{nwf}}(2n\tau)$ is the non-wavefront, which returns to the input end after several reflections. These definitions lead to the two expressions below:

$$V_{r_{wf}}(2n\tau) = \Gamma_{n,\,n+1} \cdot \prod_{j=1}^{n} \left(T_{j-1,\,j} \cdot T_{j,\,j-1}\right) \tag{7}$$

$$\begin{aligned} V_{r_{nwf}}(2n\tau) = {} & V_e(0) \cdot \Gamma_{n-2,\,n-1} \cdot \prod_{j=1}^{n-2} \left(T_{j-1,\,j} \cdot T_{j,\,j-1}\right) \\ & \cdot \sum_{j=1}^{n-2} \left\{ \left(3 - \frac{2j}{n-2}\right) \cdot \Gamma^2_{j-1,\,j} \cdot \Gamma^2_{j,\,j+1} \right. \\ & \left. + \left(2 - \frac{j}{n-2}\right) \Gamma_{j-1,\,j} \cdot \Gamma_{j,\,j+1} \cdot \sum_{k=1}^{j-1} \Gamma_{k-1,\,k} \cdot \Gamma_{k,\,k+1} \right\} \end{aligned} \tag{8}$$

with $\frac{j}{n-2} = \begin{cases} 0, & j < n-2 \\ 1, & j = n-2 \end{cases}$.

As the reflected waves are infinitive and complicated in calculating, it's just counted until the one of reflections order 5 because of the more reflection order, the smaller the value and the more complex the calculation. In the other hand, it is supposed that all the waves having more than 5 reflections are negligible.

Assuming that the input wave is a Dirac delta function ($V_e(t) = 0\,\forall t \neq 0$) to simplify the formula, and that $\Gamma_{i,\,i+1}$, $T_{i,\,i+1}$ ($i = 0, 1, \ldots, n-1$) are determined previously, just one unknown which concerns to the farthest section must be found:

$$\begin{aligned}\Gamma_{n,\,n+1} &= \frac{V_{r_{wf}}(2n\tau)}{V_e(0)\cdot\prod_{j=1}^{n}\left(T_{j-1,\,j}\cdot T_{j,\,j-1}\right)} \\ &= \frac{V_r(2n\tau)-V_{r_{nwf}}(2n\tau)}{V_e(0)\cdot\prod_{j=1}^{n}\left(T_{j-1,\,j}\cdot T_{j,\,j-1}\right)}\end{aligned} \tag{9}$$

Then the new impedance of the line is easy to be calculated by:

$$Z_{n+1} = \frac{1+\Gamma_{n,\,n+1}}{1+\Gamma_{n,\,n+1}} Z_n \tag{10}$$

As a result, the various impedance of the previous specified filter is shown in Figure 5.

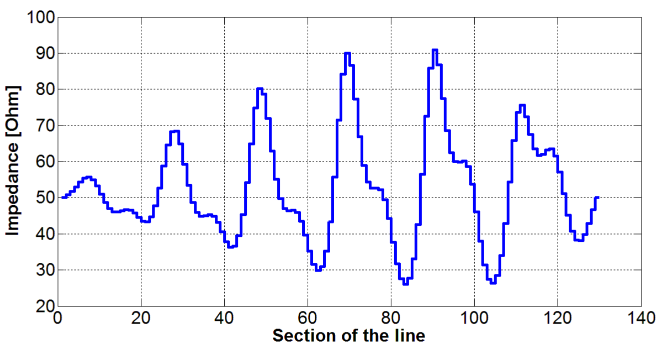

Figure 5. Impedance profile of the filter.

2.3. Step 3: Filter Dimension Calculation

The last step for designing the NTL filter profile is the calculation of the W/h ratio from known expressions [17]. It should noted that the length of line is always:

$$L = l \times N \tag{11}$$

Assuming a FR4 dielectric with $\varepsilon_r = 4,3$, the designed filter is shown in Figure 6.

3. RESULT EVALUATION

3.1. Simulation in Ideal Electric Conditions

Considering the model of each elementary line section, it is possible to calculate the transmission and reflection parameters, S_{21} and S_{11}

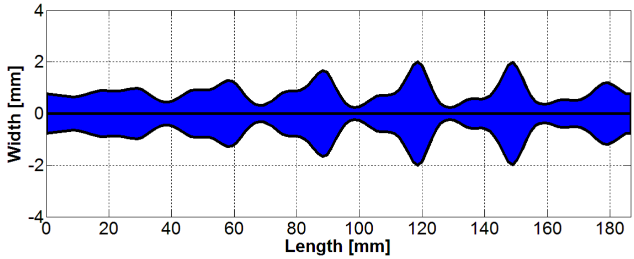

Figure 6. Filter profile.

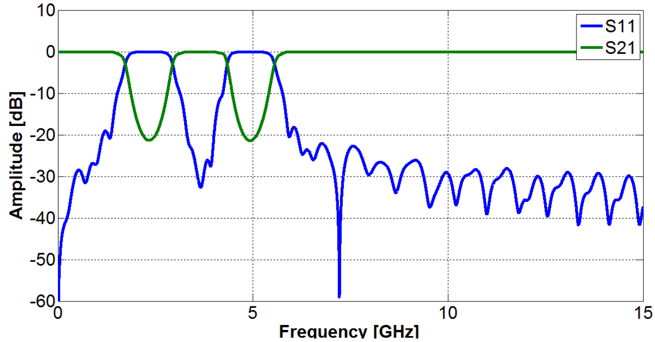

Figure 7. Obtained S parameters by electric simulation.

respectively. A simple approach is to consider the $ABCD$ matrix which models the NTL filter. This matrix is the product of $ABCD$ matrices of each elementary line because the total line can be considered as the cascade of simple lines. Afterward, the $ABCD$ matrix is converted to S parameters [19] as follow:

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \frac{1}{Z_0A + B + Z_0^2C + Z_0D} \cdot \begin{bmatrix} Z_0A+B-Z_0^2C-Z_0D & 2Z_0(AD-BC) \\ 2Z_0 & -Z_0A+B-Z_0^2C+Z_0D \end{bmatrix} \quad (12)$$

The designed filter can be then evaluated determining the S parameters. For the considered example, the results are illustrated in Figure 7. The S parameters are in good agreement compared to the filter specifications. The coefficient S_{11}, that shows the reflection loss, is less than -10 dB in the following frequency bands: [2–2.8 GHz] and [4.6–5.3 GHz]. The parameter S_{21}, that traduces the insertion loss, is less than -10 dB in [1.8–3 GHz] and [4.4–5.6 GHz] bands. It

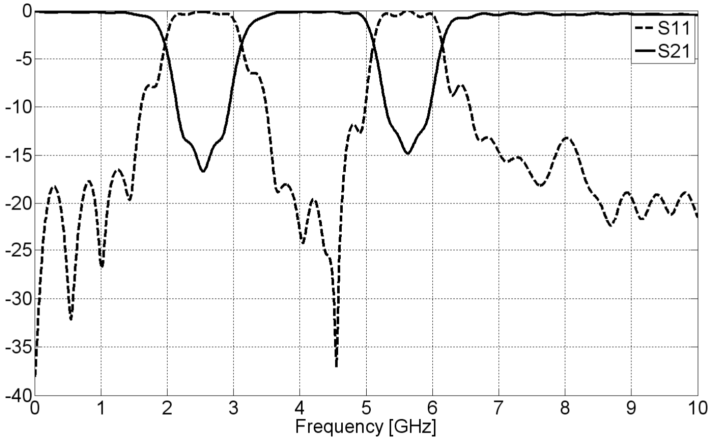

Figure 8. Obtained S parameters by electromagnetic simulation.

also presents a flat response in the passband. This first evaluation step having validated the designed profile, a second step is to realize a simulation with electromagnetic software taking into account more physical effects.

3.2. Simulation with Electromagnetic Software

A more complete evaluation is to simulate the performance of the filter with electromagnetic software. In this study, CST Microwave Studio has been used. This evaluation step can be realized quickly by importing the filter profile in format ".dxf" file. The simulation result is showed in Figure 8.

In Figure 8, the filter specification has the same profile as it is simulated in Matlab but with slight ripples in the passband and a little frequency-shift. This last phenomenon is caused by the energy storage at the discontinuities of the line, and one of the solutions for compensating is proposed [16].

3.3. Constraints and Limitations

3.3.1. Incertitude Due to Approximations

The approximations in the process (V_r discrete, reflected order, etc.) decrease the gain and loss of desired filter as illustrated for example in Figure 9. To compensate this loss, the gain of S parameter can be set positive. However, since it is a passive filter, the result of gain obtained will be always 0 dB maximum.

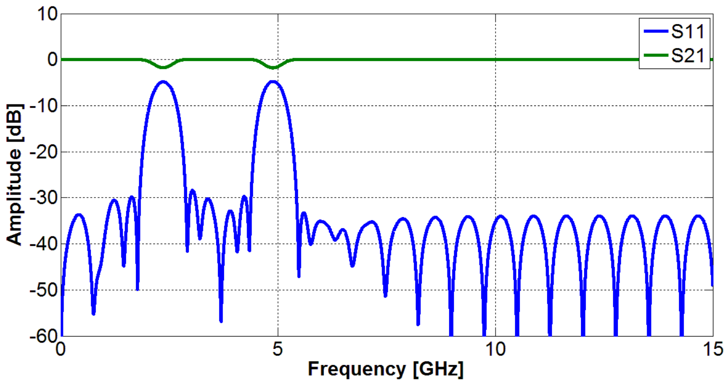

Figure 9. The S parameters without compensation.

3.3.2. Input and Output Impedance

Almost the radio frequency components are adapted to standard impedance $50\,\Omega$ at input and output. According to (7) and (8), when V_r gets the steady-state at the time '($V_r = constant, \forall t > t$))', then $Z_n = Z_{n+1} = \ldots = Z_{filter\_output}$. In case the steady-state value of V_r is 0, it's showed up again the ideal condition that $Z_{filter_{output}} = Z_{filter_{intput}} = 50\,\Omega$ as default.

3.3.3. Reflected Energy

One of the inconveniences of this method (and the associated technology) is the technical limitation concerning to the printable strip on the electronic boards. That means the width of the line couldn't be so thin (Z_i so grand) or so large (Z_i so small). Figure 10 illustrates this remark with an example ($Z_{\max} \geq 150\,\Omega$).

Otherwise, it is certainly impractical in case of the small impedance which gives a very large strip on the electric board.

In the other hand, the value of impedance Z depends on the reflected wave V_r, so $Z_{\max}$ depends on $V_{r\_\max}$ which is decided by the positive gain and the rejected bandwidth of the S_{11} parameters, called the reflected energy.

Since the reflected energy of a lowpass and bandpass filter is so high and due to the impedance limitation, it is so difficult to create such these filters by this method, but a highpass, stopband and multiband filters are perfectly appropriate.

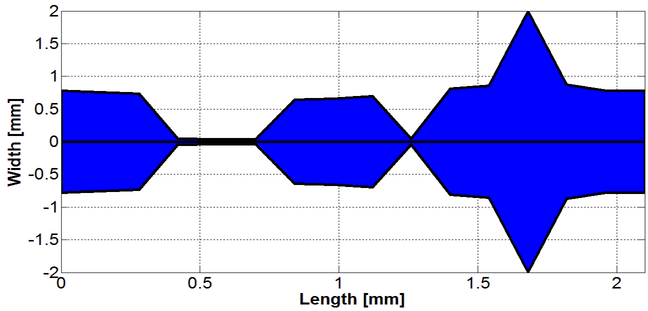

Figure 10. Example of a filter profile with the microstrip of 150 Ω (0.08 mm).

3.3.4. The Filter Length

Certainly, the filter length relies on τ and ν, and ν depends on the material dielectric coefficient ε_r. Hence, to reduce the length (due to some technical limitations), it is possible to change the material (i.e., change ε_r) so that the length will also be changed. Besides, since the filter specification relies only on its impedance not on the length of the filter (as showed in the result evaluation part), the change of filter length will optimize the filter performance.

4. CONCLUSION

A method to design filters based on the nonuniform transmission line is presented. This method promises an advanced technique in filter design domain thanks to its simplicity and its totally automatic process. In addition, the type of designed UWB multiband filter offers new relevant solutions in UWB context.

REFERENCES

1. Matthaei, G., L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-matching Networks, and Coupling Structures*, Artech House, 1980.
2. Huang, S. Y. and Y. H. Lee, "Development of Ultra-Wideband (UWB) filters," *General Assembly of the International Union of Radio Science*, Chicago, Illinois, USA, August 7–16, 2008.
3. Zhu, L., S. Sun, and W. Menzel, "Ultra-Wideband (UWB) bandpass filters using multiple-mode resonator," *IEEE Microwave*

and Wireless Components Letter, Vol. 15, No. 11, 796–798, November 2005.
4. Li, R. and L. Zhu, "Compact UWB bandpass filter using stub-loaded multiple-mode resonator," *IEEE Microwave and Wireless Components Letters*, Vol. 17, No. 1, 40–42, January 2007.
5. Yao, B., Y. Zhou, Q. Cao, and Y. Chen, "Compact UWB bandpass filter with improved upper-stopband performance," *IEEE Microwave and Wireless Components Letters*, Vol. 19, No. 1, 27–29, January 2009.
6. Cai, P., Z. Ma, X. Guan, Y. Kobayashi, T. Anada, and G. Hagiwara, "A compact UWB bandpass filter using two section open-circuited stubs to realize transmission zeros," *Proc. of Asia-Pacific Microwave Conference*, 3356–3359, Suzhou, China, December 4–7, 2005.
7. Hsu, C. L., F. C. Hsu, and J. T. Kuo, "Microstrip bandpass filters for Ultra-Wideband (UWB) wireless communications," *Proc. of International Microwave Symposium*, 679–682, Long Beach, CA, USA, June 15–17, 2005.
8. Chen, N. W. and K. Z. Fang, "An ultra-broadband coplanar-waveguide bandpass filter with skirt selectivity," *IEEE Microwave and Wireless Components Letters*, Vol. 17, No. 2, 124–126, February 2007.
9. Wang, H., L. Zhu, and W. Menzel, "Ultra-wideband bandpass filter with hybrid microstrip/CPW structure," *IEEE Microwave and Wireless Components Letters*, Vol. 15, No. 12, 844–846, December 2005.
10. Sanchez-Soriano, M. A., E. Bronchalo, and G. Torregrosa-Penalva, "Compact UWB bandpass filter on signal interference technique," *IEEE Microwave and Wireless Components Letters*, Vol. 19, No. 11, 692–694, November 2009.
11. Ma, Z., C. P. Chen, and T. Anada, "New progress in the development of compact high-performance microwave Ultra-Wideband (UWB) filters," *Proc. of IEEE International Symposium on Radio-Frequency Integration Technology*, 355–360, Singapore, December 9–11, 2009.
12. Duroc, Y., I. Kharrat, and D. Kaddour, "Time domain design for UWB filters," *Microwave and Optical Technology Letters*, Vol. 53, No. 5, 1006–1011, May 2011.
13. Hsue, C. W. and T. W. Pan, "Reconstruction of nonuniform transmission lines from time-domain reflectometry," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 45, No. 1, 32–38, January 1997.

14. Pan, T. W. and C. W. Hsue, "Modified transmission and reflection coefficients of nonuniform transmission lines and their applications," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 46, No. 12, 2092–2097, December 1998.
15. Roy, M. L., A. Pérennec, S. Toutain, and L. C. Calvez, "The continuously varying transmission-line technique — Application to filter design," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 47, No. 9, 1680–1687, September 1999.
16. Nakayama, M., H. Deguchi, M. Tsuji, and H. Shigesawa, "Design of microstrip-line bandpass filter based on reflected characteristics in time domain," *Electronics and Communications in Japan*, Part 2, Vol. 89, No. 3, 1–9, March 2006.
17. Pan, T. W., C. W. Hsue, and J. F. Huang, "Time-domain reflectometry using arbitrary incident waveforms," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 50, No. 11, 2558–2263, November 2002.
18. Isydorczyk, J., "Comments on time-domain reflectometry using arbitrary incident waveforms," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 51, No. 4, 1296–1297, April 2003.
19. Pozar, D. M., *Microwave Engineering*, 3rd edition, Wiley, 2005.