

# Dual Band Wilkinson Power Dividers Using T-Sections

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**Abstract**— This paper presents analysis and design of a dual band Wilkinson power divider for wireless applications. The design is accomplished by transforming the lengths and impedances of the quarter wavelength sections of the conventional Wilkinson power divider into dual band T-shaped sections. Simple design equations for the basic geometry are given. The designed dual band Wilkinson power divider structure is analyzed using a full wave electromagnetic EM simulator and then realized for the Global system of mobile communications (GSM) bands (925-960 MHz) and (1930-1960 MHz). The computed and the measured S-parameters are in a very good agreement.

**Index Terms**— Dual band, Microstrip, T-section, Wilkinson power divider.

## I. INTRODUCTION

The Wilkinson power dividers and combiners are among the most used components in wireless communication systems for power division or combination. They are widely used in microwave power amplifier linearization, high power transmitter, antenna array feed networks and in calibration laboratories. These power dividers generally employ quarter wave length transmission line sections at the design center frequency [1] as shown in Fig. 1(a). The new technology for mobile communications led researchers to develop more versatile components that have to address the requirements relative to multi-frequency bands. Monzon [2] illustrated the extension of the two section transformer to operate at two arbitrary frequencies. A Wilkinson power divider that operates at one design frequency  $f_1$  and its first harmonics ( $2f_1$ ) was presented in [3]. A Wilkinson power divider which operates at any two arbitrary frequencies of interest  $f_1$  and  $mf_1$  (where  $m$  is the frequency ratio  $m > 1$ ) was suggested in [4], in which, each quarter wavelength branch of the conventional Wilkinson power divider is substituted by two sections of wave transmission line and lumped parameters (R, L and C) connected between the two output ports as depicted in Fig. 1(b). In [5], the conversion of the quarter wavelength sections of the conventional power divider into their dual band II-section equivalents is used. The dual band divider designed using this method is operating at 1.8 and 4.775 GHz with a frequency ratio of 2.65. Despite its reduced size, this divider can be only, practically, designed with a high frequency ratio; on the range of (2.5 to 5.5). This disadvantage make the design not suitable for dual band operation with lower frequency ratios that is

needed for systems such as GSM with its dual bands specified at (925-960MHz) and (1930-1960MHz).

In this paper, we propose a new configuration for dual band Wilkinson power dividers that can be used at the receiving bands of the Global system for mobile communication GSM (925-960 MHz) and GSM (1930-1960 MHz) with only one lumped resistance element (R). This dual band Wilkinson power divider is design by converting each quarter wavelength section of the conventional power divider of Fig. 1(a) to its equivalent T-shaped transmission line section. The details of the procedures are described in the following sections.

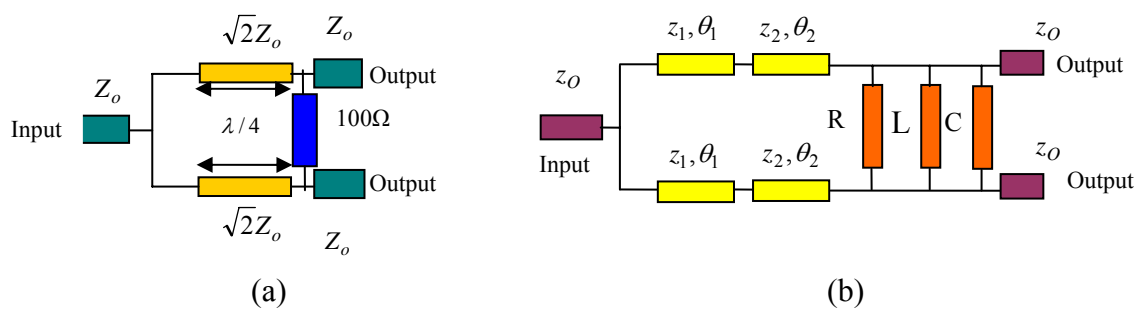


Fig. 1. The Wilkinson power divider:

(a) The traditional power divider and (b) The 2-section dual-band power divider.

## II. DUAL BAND WILKINSON POWER DIVIDER USING T-SHAPED TRANSMISSION LINE

Each quarter wave length section of the conventional Wilkinson power divider shown in Fig.1a will be converted to a dual band T-shaped transmission line section. The equivalence between the impedance of a transmission line section shown in Fig. 2(a), and the T-shaped section of Fig. 2(b) is investigated utilizing the ABCD matrices of both sections.

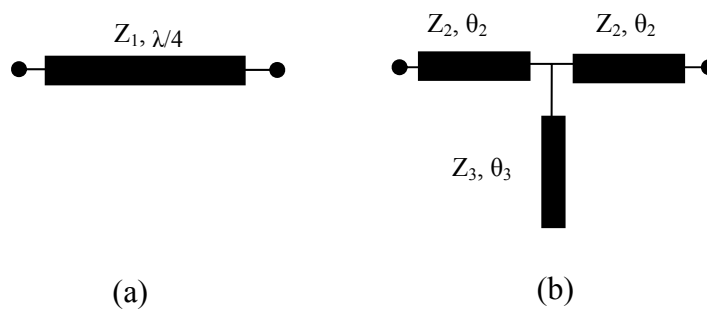


Fig. 2. A quarter-wave transmission line (a) and its T-equivalent (b).

The ABCD matrix for the original transmission line, which is  $(\lambda_g/4)$  in length shown in Fig. 2(a), will be [6]:

$$M_1 = \begin{bmatrix} 0 & jZ_1 \\ jY_1 & 0 \end{bmatrix} \tag{1}$$

The ABCD matrix for the T-shaped transmission line section shown in Fig. 2(b) is

$$M_T = M_2 M_3 M_2 \tag{2a}$$

where  $M_2$  and  $M_3$  are the ABCD matrices for the series and shunt elements and are defined as follows:

$$M_2 = \begin{bmatrix} \cos \theta_2 & jZ_2 \sin \theta_2 \\ jY_2 \sin \theta_2 & \cos \theta_2 \end{bmatrix} \tag{2b}$$

$$M_3 = \begin{bmatrix} 1 & 0 \\ jY_3 \tan \theta_3 & 1 \end{bmatrix} \tag{2c}$$

By equating the elements A (of the ABCD matrix) from both Eq. (1) and Eq. (2), we get:

$$\tan \theta_3 = 2 \frac{Z_3}{Z_2} \cot(2\theta_2) \tag{3}$$

Equating the element B (of the ABCD matrix) from Eqs (1) and (2) results in:

$$Z_1 = 2Z_2 \sin \theta_2 \cos \theta_2 - Z_2^2 Y_3 \sin^2(\theta_2) \tan \theta_3 \tag{4}$$

Substituting Eq. (3) into Eq. (4) results in

$$Z_1 = Z_2 \tan \theta_2 \tag{5}$$

For the purpose of dual band operation, it is necessary to modify Eq. (5) as follows [7]:

$$Z_2 \tan \theta_{2(f_1)} = \pm Z_1 \tag{6a}$$

$$Z_2 \tan \theta_{2(f_2)} = \pm Z_1 \tag{6b}$$

Where  $\theta_{2(f_1)}$  and  $\theta_{2(f_2)}$  are the electrical lengths of the series element of the stub at the two operating frequencies  $f_1$  and  $f_2$  respectively and ( $f_2 > f_1$ ). The solution for Eq. (6) is given by:

$$\theta_{2(f_2)} = n\pi + \theta_{2(f_1)} \quad n=1, 2, \dots \tag{7}$$

$$\theta_{2(f_2)} / \theta_{2(f_1)} = f_2 / f_1 = R \tag{8}$$

Based on Eq. (3) and Eq. (8), the following condition can be derived:

$$\theta_{3(f_2)} = m\pi \pm \theta_{3(f_1)} \quad m=1, 2, \dots \tag{9}$$

$$\theta_{3(f_2)} / \theta_{3(f_1)} = f_2 / f_1 = R \tag{10}$$

For a compact size transmission line,  $n = m = 1$  must be selected. From the above equations, the practical electrical lengths of the T-shape sections must be bounded in these regions ( $\pi/4 < \theta_2 < \pi/2$ ) and ( $\pi/2 < \theta_3 < \pi$ ). The design procedure for power divider coupler that operates at dual band can be summarized as follows:

-The value  $R$  is found from Eq. (8)

-The electrical length corresponding to series transmission line will be:

$$\theta_{2(f_1)} = \pi / (R + 1) \tag{11a}$$

$$\theta_{2(f_2)} = R \cdot \theta_{2(f_1)} \tag{11b}$$

-The electrical length corresponding to shunt stub will be:

$$\theta_{3(f_1)} = 2\pi / (R + 1) = 2\theta_{2(f_1)} \tag{12a}$$

$$\theta_{3(f_2)} = R \cdot \theta_{3(f_1)} = 2\theta_{2(f_2)} \tag{12b}$$

-The values of impedances  $z_2$  and  $z_3$  can be calculated as follows:

$$Z_2 = Z_1 / \tan \theta_{2(f_1)} \tag{13a}$$

$$Z_3 = 0.5Z_2 \tan^2(\theta_{3(f_1)}) \tag{13b}$$

Fig. 3 illustrates the variations of the electrical lengths  $\theta_{2(f_1)}$  and  $\theta_{3(f_1)}$  (in degrees) against the frequency ratio ( $R$ ), while Fig. 4 illustrates the variation of the characteristic impedances  $Z_2$  and  $Z_3$  against the frequency ratio ( $R$ ).

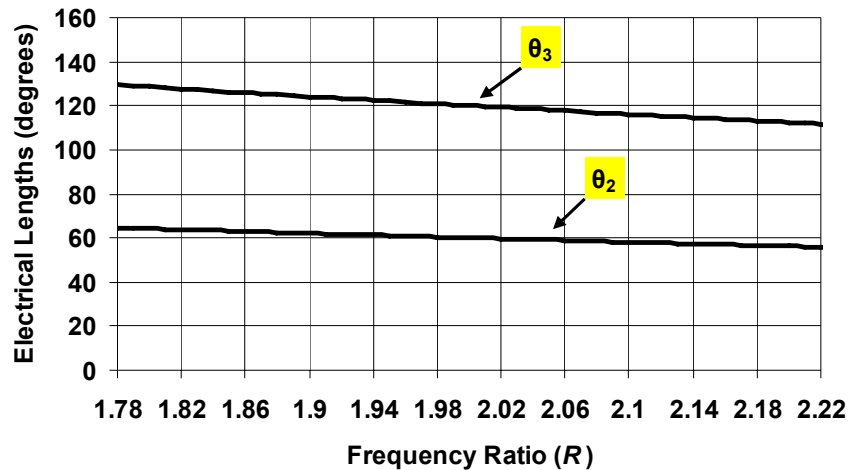


Fig. 3. The variation of electrical lengths against the frequency ratio ( $R$ ).

### III. DESIGN, SIMULATION, REALIZATION AND MEASUREMENTS

A dual band Wilkinson power divider is designed to operate in the GSM-bands (925-960MHz) and (1930-1960MHz). The required operating frequencies are 942.5 MHz and 1945 MHz and the frequency ratio will be  $R=2.064$ . Each  $\lambda/4$  branch of the conventional Wilkinson power divider of Fig. 1(a) is replaced by its corresponding dual band T-section. The values for  $\theta_{2(f_1)}$  and  $\theta_{3(f_1)}$ ,  $Z_2$  and  $Z_3$  will be  $58.75^\circ$ ,  $117.51^\circ$ ,  $42.91 \Omega$  and  $79.102 \Omega$ , respectively. The configuration, shown in Fig. 5(a) has been investigated using a full wave EM simulator, and then realized on RT/Duroid 5880 ( $\epsilon_r=2.2$ ,  $h=1.57\text{mm}$ ). The realized divider is shown in Fig. 5(b). Unlike the structure presented in [5], the proposed configuration has only one open stub and is suitable for lower frequency ratio dual band operation as it is the case in modern wireless systems.

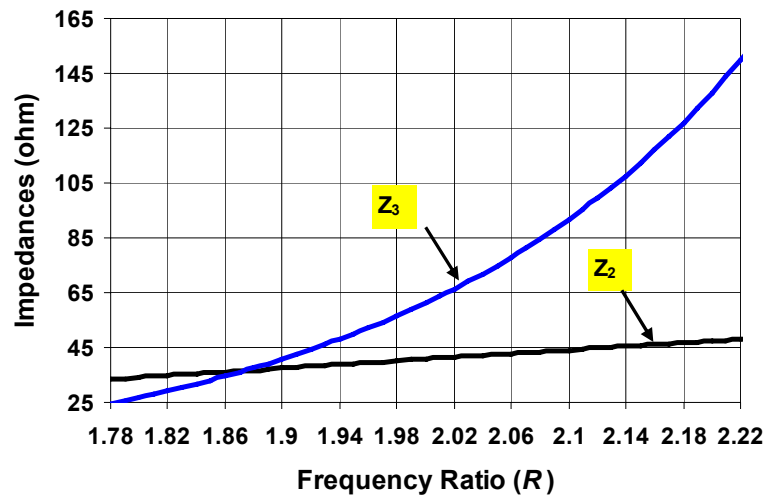


Fig. 4. The impedance variations against the frequency ratio ( $R$ ).

The results of the simulation and measurements are shown in Figs. 6-9. The measured (M) return loss  $S_{11}$  is less than (-24 dB) for both operating GSM-bands (925-960 MHz) and (1930-1960 MHz). The measured (M) insertion parameter  $S_{21}$  is quite around (-3 dB) for both frequency bands with a deviation of (0.3 dB) for the first frequency band and (0.4 dB) for the second frequency band. The measured insertion parameter ( $S_{31}$ ) is around (-3 dB) for both frequency bands and it deviates only by (0.35 dB) for the first frequency band and (0.4 dB) for the second frequency band. The measured Return loss ( $S_{22}$ ) is less than (-27 dB) for the first band and less than (-28 dB) for the second operating band. The isolation coefficient ( $S_{23}$ ) is less than (-25 dB) in the first frequency band and less than (-23 dB) in the second frequency band. The measured (M) phase shift between the output ports does not exceed ( $3^\circ$ ) for the first frequency band and is less than ( $6^\circ$ ) for the second frequency band.

#### IV. CONCLUSIONS

This paper presents analysis and design of a dual band Wilkinson power divider. The design uses dual band T-section transformation of the quarter wavelength sections of the divider without any short circuit stubs or via holes. Practical and simple design equations are presented. A dual band GSM Wilkinson power divider is designed and examined using a full wave EM software simulator and then realized on RT/Duroid 5880 substrate ( $\epsilon_r=2.2$ ,  $h=1.57$  mm). The measurements of the realized dual-band Wilkinson power divider are in excellent agreement with the simulated results. The coupling coefficients ( $S_{21}$ ) and ( $S_{31}$ ) are around (-3 dB) with variation less than (0.4 dB) for both of the operating GSM bands. The output phase shift variation does not exceed ( $3^\circ$ ) for the first frequency band and it is less than ( $6^\circ$ ) for the second frequency band.

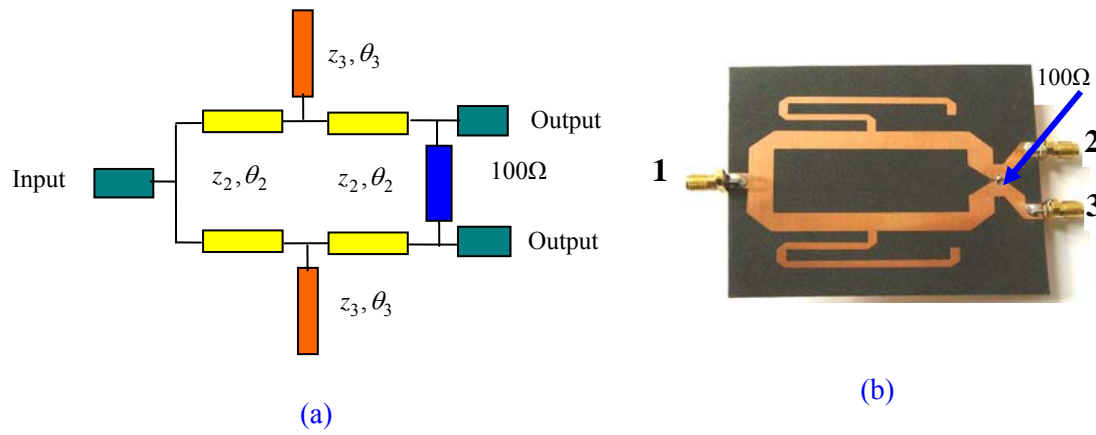


Fig. 5. The dual band power divider using T-sections:  
 (a) The divider schematic and (b) The realized divider.

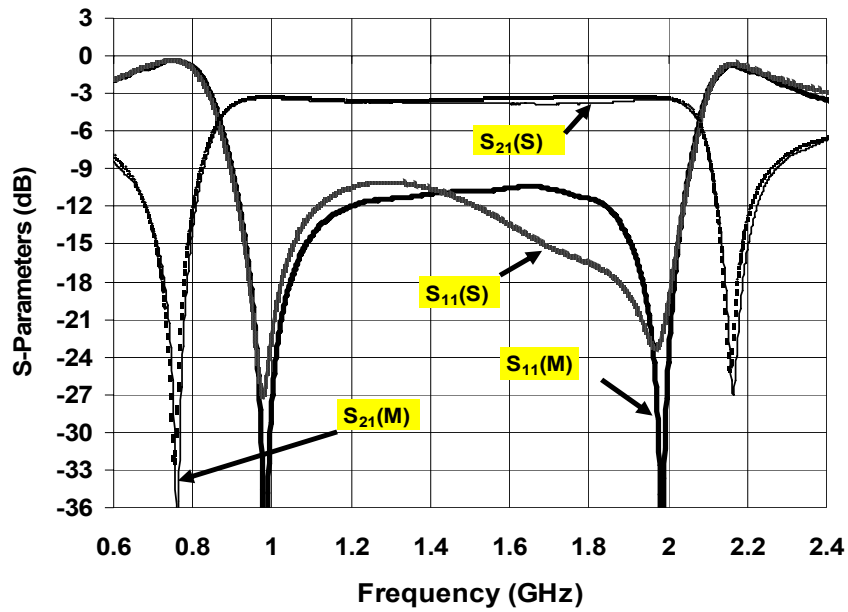


Fig. 6. The measured (M) and the simulated (S) return loss ( $S_{11}$ ) and insertion loss ( $S_{21}$ ) against frequency.

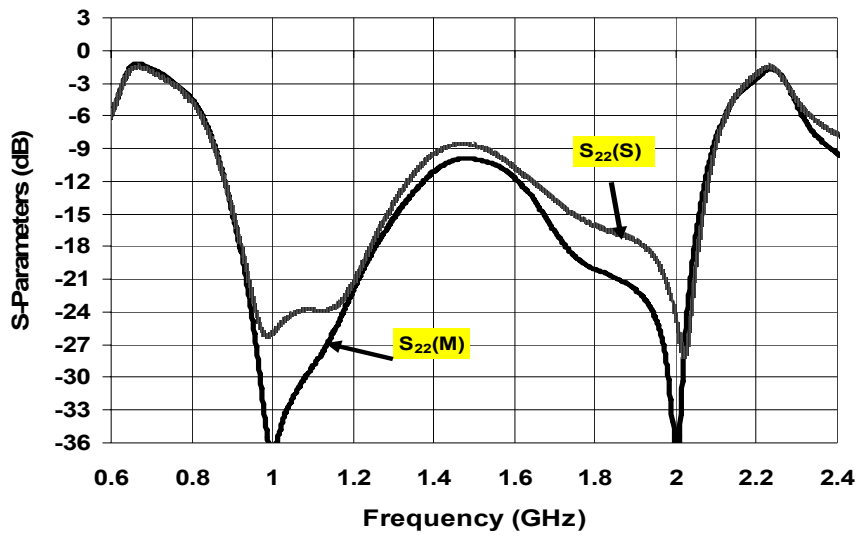


Fig. 7. The measured (M) and the simulated (S) return loss ( $S_{22}$ ) against frequency.

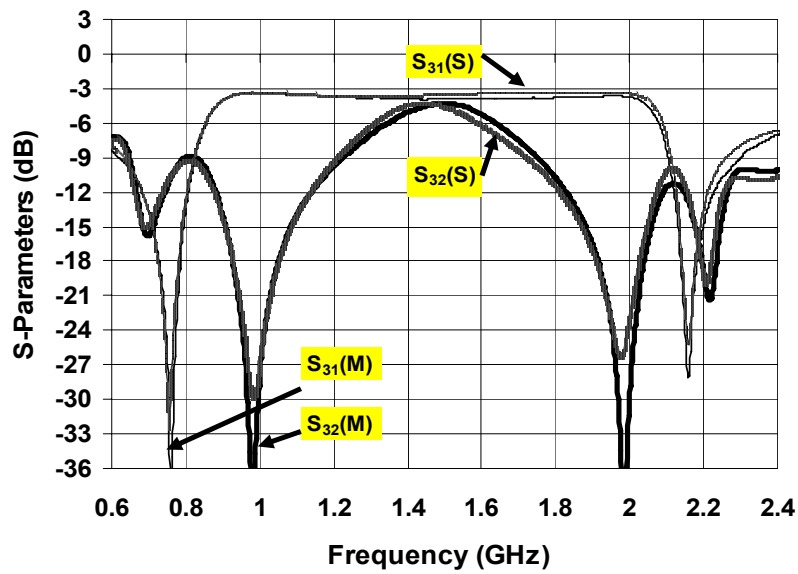


Fig. 8. The measured (M) and the simulated (S) isolation loss ( $S_{32}$ ) and insertion loss ( $S_{31}$ ) against frequency.

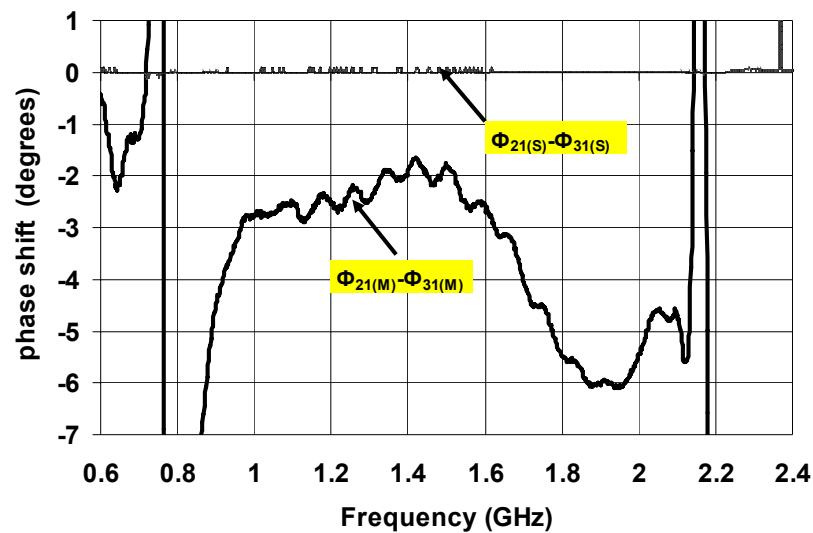


Fig. 9. The measured (M) and the simulated (S) phase shift between the output two ports.

#### ACKNOWLEDGMENT

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#### REFERENCES

- [1] E. J. Wilkinson, "An N-way hybrid power divider," *IRE Trans. Microwave Theory Tech.*, vol. MTT-8, no. 1, pp. 116-118, Jan 1960.
- [2] Cesar Monzon, "A small dual-frequency transformer in two sections," *IEEE Trans. Microwave and Theory Tech.*, vol. 51, no. 4, pp. 1157-1161, April 2003.
- [3] Lei Wu, T. Bitzer and A. Berroth, "A dual-frequency Wilkinson power divider: for a frequency and its first harmonics," *IEEE Trans. Microwave and Wireless Components Letters*, vol. 15, no. 2, pp. 107-109, Feb. 2005.
- [4] Lei Wu, Z. Sun, H. Yilmaz and M. Berroth, "A dual-frequency Wilkinson power divider," *IEEE Trans. Microwave and Theory Tech.*, vol. 54, no. 1, pp. 278-284, Jan. 2006.
- [5] Ashraf S. Mohra, "Compact Dual band Wilkinson power divider," *Microwave and optical technology letters*, vol. 50, no. 6, pp. 1678-1682. June 2008.
- [6] David M. Pozar, " *Microwave Engineering*," John Wiley & Sons Inc., 3<sup>rd</sup> edition, 2005.
- [7] Kwok Keung, M. Cheng and Fai -Leung. Wong, "A novel approach to the design and implementation of dual band compact planar 90<sup>o</sup> branch line coupler," *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 11, pp. 2458-2463, Nov. 2004.