Short Papers

Coaxial Impedance Inverter Design

A. GIEFING

***Abstract*—It is shown that the effects of discontinuity capacitances associated with the abrupt change in the center conductor diameter of a coaxial bandpass filter can readily be anticipated at the designing stage.**

Recently Davis and Khan [1] published a paper on "coaxial bandpass filter design," based on an improved design of an impedance inverter with the distributed line lengths of the inverter taken into account. The line elements obtained in this manner can be corrected for discontinuity capacitances by any of several methods.

It will be shown that this can be avoided if the reactive elements resulting from the discontinuity are taken into account from the outset.

We start by calculating the transfer matrix of the circuit shown in Fig. 1, consisting of a line length l with characteristic impedance Z_0 together with two capacitances C_d and two lengths $\Phi/2$ with the characteristic impedance Z_c, all assumed to be loss-free.

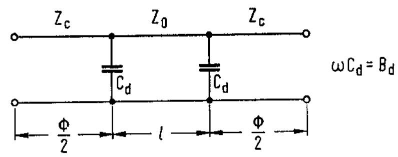

Fig. 1. Equivalent circuit for a disk impedance inverter.

$$\left.\begin{aligned}
A = D = &-\left[B_d Z_0 \cos\Phi + \frac{1}{2}\left(\frac{Z_c}{Z_0} + \frac{Z_0}{Z_c} - B_d^2 Z_0 Z_c\right)\sin\Phi\right] \\
&\cdot \sin\beta l + [\cos\Phi - B_d Z_c \sin\Phi]\cos\beta l \\
j\frac{B}{Z_c} = &\left[2B_d Z_0 \sin\frac{\Phi}{2}\cos\frac{\Phi}{2} - \frac{Z_0}{Z_c}\cos^2\frac{\Phi}{2}\right. \\
&\left. - \left(B_d^2 Z_0 Z_c - \frac{Z_c}{Z_0}\right)\sin^2\frac{\Phi}{2}\right]\sin\beta l \\
&+ \left[-2\sin\frac{\Phi}{2}\cos\frac{\Phi}{2} + 2B_d Z_c \sin^2\frac{\Phi}{2}\right]\cos\beta l \\
-jCZ_c = &\left[-2B_d Z_0 \sin\frac{\Phi}{2}\cos\frac{\Phi}{2} - \frac{Z_0}{Z_c}\sin^2\frac{\Phi}{2}\right. \\
&\left. - \left(B_d^2 Z_0 Z_c - \frac{Z_c}{Z_0}\right)\cos^2\frac{\Phi}{2}\right]\sin\beta l \\
&+ \left[2\sin\frac{\Phi}{2}\cos\frac{\Phi}{2} + 2B_d Z_c \cos^2\frac{\Phi}{2}\right]\cos\beta l
\end{aligned}\right\} . \tag{1}$$

Combining the last two equations of (1) gives

$$j\left(\frac{B}{Z_c} - CZ_c\right) = \left(\frac{Z_c}{Z_0} - \frac{Z_0}{Z_c} - B_d^2 Z_0 Z_c\right)\sin\beta l + 2B_d Z_c \cos\beta l. \tag{2}$$

To obtain the immittance inverting operation at a certain frequency the elements of the above matrix must be chosen so that

$$\left.\begin{aligned}
A &= D = 0 \\
B &= jK \\
C &= \frac{j}{K}
\end{aligned}\right\} \tag{3}$$

where K is a real constant.

From the last two equations of (3) it follows that

$$j\left(\frac{B}{Z_c} - CZ_c\right) = \frac{Z_c}{K} - \frac{K}{Z_c}. \tag{4}$$

Equating the terms on the right-hand sides of (2) and (4) we now find

$$\left(\frac{Z_c}{Z_0} - \frac{Z_0}{Z_c} - B_d^2 Z_0 Z_c\right)\sin\beta l + 2B_d Z_c \cos\beta l = \frac{Z_c}{K} - \frac{K}{Z_c}.$$

Reduced to a quadratic equation of $\tan\beta l$, we obtain for the unknown length l

$$\tan^2\beta l + \frac{2ab}{a^2 - c^2}\tan\beta l + \frac{b^2 - c^2}{a^2 - c^2} = 0 \tag{5}$$

with

$$\left.\begin{aligned}
a &= \frac{Z_c}{Z_0} - \frac{Z_0}{Z_c} - B_d^2 Z_0 Z_c \\
b &= 2B_d Z_c \\
c &= \frac{Z_c}{K} - \frac{K}{Z_c}
\end{aligned}\right\} . \tag{6}$$

Now knowing $\tan\beta l$, from the condition $A = D = 0$ of (3) $\tan\Phi$ may be found from (1)

$$\tan\Phi = \frac{1 - B_d Z_0 \tan\beta l}{B_d Z_c + \frac{1}{2}\left(\frac{Z_0}{Z_c} + \frac{Z_c}{Z_0} - B_d^2 Z_0 Z_c\right)\tan\beta l}. \tag{7}$$

References

[1] W. A. Davis and P. J. Khan, "Coaxial bandpass filter design," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 373–380, Apr. 1971.

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The author is with Siemens AG, Wv ZL Laboratory 545, Munich 70, Germany.

Comments on "A Relationship Between the Scattering Parameters of a Passive Lossy Nonreciprocal Two-Port"

JEROME I. GLASER

***Abstract*—An inequality relating the scattering coefficients of a passive lossy-reciprocal or nonreciprocal two-port is derived. For the reciprocal two-port the inequality reduces to that presented by Uhlir.**

Introduction

In this letter we derive an upper-bound inequality that applies to the scattering coefficients of a passive lossy nonreciprocal two-port. The inequality given here corrects the inequality derived in the above correspondence,[1] and reduces to the inequality given by Uhlir [1] for the reciprocal case. This inequality allows one to bound the magnitude of the reflection coefficient of one port using measurements of the magnitudes of the reflection coefficient of the other port and the insertion losses of the network in both directions. Such a bound simplifies the automated testing of microwave two-ports.

Let the elements of the scattering matrix of a two-port be given by S_{11}, S_{12}, S_{21}, S_{22} and their phases be given by θ_{11}, θ_{12}, θ_{21}, and θ_{22}, respectively.

Equation (3) of footnote one is given by:

$$(1 - |S_{11}|^2 - |S_{21}|^2)(1 - |S_{22}|^2 - |S_{12}|^2) \geq |S_{11}^* S_{12} + S_{21}^* S_{22}|^2 \tag{1}$$

where $|\ |$ denotes magnitude and S^* denotes the complex conjugate of S. After putting only those terms that contain $|S_{22}|$ on the right-

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The author is with Bell Telephone Laboratories, Inc., Murray Hill, N. J. 07974.
[1] H. J. Hindin, *IEEE Trans. Microwave Theory Tech.* (Corresp.), vol. MTT-19, p. 781, Sept. 1971.

hand side of the inequality, (1) can be rearranged to yield the form

$$(1-|S_{21}|^2)(1-|S_{12}|^2)-|S_{11}|^2 \geq |S_{22}|^2(1-|S_{11}|^2) + 2|S_{22}||S_{11}||S_{12}||S_{21}| \cos\theta \quad (2)$$

where

$$\theta = \theta_{11} + \theta_{22} - \theta_{21} - \theta_{12}. \quad (3)$$

Dividing both sides of (2) by the factor $(1-|S_{11}|^2)$ and then completing the square on the right-hand side gives the following equation:

$$\frac{(1-|S_{21}|^2)(1-|S_{12}|^2)-|S_{11}|^2}{1-|S_{11}|^2} + \frac{|S_{11}|^2|S_{21}|^2|S_{12}|^2\cos^2\theta}{(1-|S_{11}|^2)^2} \geq \left(|S_{22}| + \frac{|S_{11}||S_{21}||S_{12}|\cos\theta}{(1-|S_{11}|^2)}\right)^2. \quad (4)$$

After taking the positive square root of (4) and rearranging terms we obtain

$$\sqrt{\frac{(1-|S_{21}|^2)(1-|S_{12}|^2)-|S_{11}|^2}{1-|S_{11}|^2} + \frac{|S_{11}|^2|S_{21}|^2|S_{12}|^2\cos^2\theta}{(1-|S_{11}|^2)^2}} - \frac{|S_{11}||S_{21}||S_{12}|\cos\theta}{1-|S_{11}|^2} \geq |S_{22}|. \quad (5)$$

The upper bound for (5) occurs for $\theta=\pi$ and is

$$\frac{\sqrt{(1-|S_{21}|^2-|S_{11}|^2)(1-|S_{12}|^2-|S_{11}|^2)} + |S_{11}||S_{21}||S_{12}|}{1-|S_{11}|^2} \geq |S_{22}|. \quad (6)$$

Equation (6) differs from eq. (9) of Hindin by the sign in front of the square-root term.

Equation (9) in Hindin leads to a contradiction as can be seen by setting $|S_{11}|=0$. In this case it gives the result

$$|S_{22}| \leq -\sqrt{(1-|S_{12}|^2)(1-|S_{21}|^2)} \quad (7)$$

which is contradictory. For $|S_{11}|=0$, (6) of this letter gives the result

$$|S_{22}| \leq \sqrt{(1-|S_{12}|^2)(1-|S_{21}|^2)} \quad (8)$$

which is not contradictory.

For a reciprocal network, (6) of this letter reduces to the form

$$\frac{|(1-|S_{12}|^2)-|S_{11}|^2| + |S_{11}||S_{12}|^2}{1-|S_{11}|^2} \geq |S_{22}|. \quad (9)$$

Using the inequality $1-|S_{12}|^2-|S_{11}|^2 \geq 0$ given in (1), (9) of this letter reduces to the form

$$1 - \frac{|S_{12}|^2}{1+|S_{11}|} \geq |S_{22}| \quad (10)$$

which is Uhlir's eq. (12).

References

[1] A. Uhlir, Jr., "Bounds on output VSWR of a passive reciprocal two-port from forward measurements," *IEEE Trans. Microwave Theory Tech.* (Corresp.), vol. MTT-18, pp. 662–663, Sept. 1970.

MIC *Ku*-Band Upconverter

P. BURA

***Abstract*—An MIC *S*- to *Ku*-band upper-sideband upconverter has shown a pump efficiency of 25 percent and 3.8-dB signal gain. When used as a lower-sideband upconverter, gains of 13 dB and 100-MHz bandwidth were measured.**

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The author was with RCA, Defense Electronic Products, Defense Communications Systems Division, Route 202, Somerville, N. J. 08876. He is now with RCA Ltd., St. Anne-de-Bellevue, Quebec, Canada.

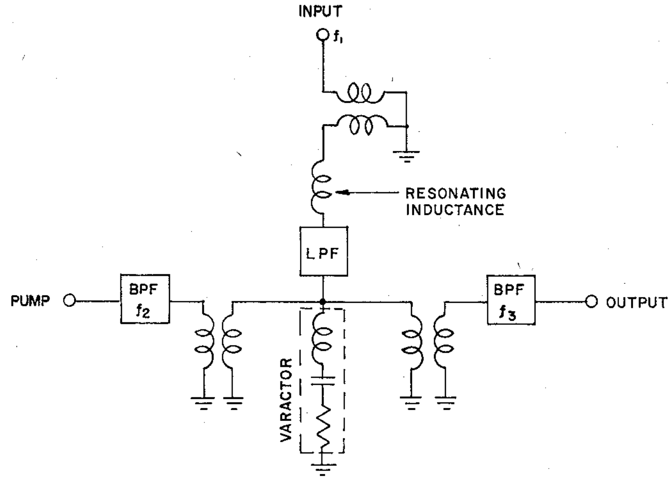

Fig. 1. Equivalent circuit of the upconverter.

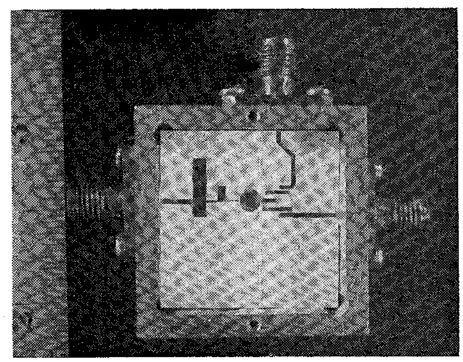

Fig. 2. MIC upconverter.

The development of the MIC *Ku*-band upconverter represents frequency extension of the previously reported *S*-band circuit [1]. The equivalent circuit of the unbalanced upconverter is shown in Fig. 1. It consists of the varactor in parallel with the input, output, and pump frequency circuits. Each circuit transforms the 50-Ω line impedance to the value required for the optimum upconverter performance and, at the same time, presents an open circuit at the other two frequencies.

The MIC upconverter, using Cr–Au metallization on a 25-mil alumina substrate, is shown in Fig. 2. The varactor was chosen to be self-resonant between the pump and the output frequencies, so that no additional tuning was required at those frequencies. Asymmetrical coupled transmission-line filters were used both for filtering and impedance transformation from $R_0=50$ Ω. The transformation is given by

$$R = \left(\frac{Z_{e2}-Z_{02}}{Z_{e1}-Z_{01}}\right)^2 R_0.$$

The values of the odd and even impedances determine the strip widths and spacings in the filters.

The signal input circuit has to resonate the varactor and to provide the correct input loading. This is achieved by means of a capacitive stub and a transforming section of a high impedance line. In addition, the input circuit contains a wide-band choke section and a pump reject stub. The upconverter design is based on the analysis of Penfield [2] and Grayzel [3].

To assure strictly reactive behavior, a constraint on the junction voltage is imposed, limiting the swing between the forward conduction and reverse breakdown. This is expressed by:

$$m_1 + m_2 + m_3 \leq 0.25$$

where m_1, m_2, and m_3 are elastance modulation ratios at component