Design of an Electronically Tunable Microwave Impedance Transformer

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Abstract **- A new technique is introduced for designing continuously variable electronically tunable microwave impedance transformers that can be implemented using MIC or MMIC technology. An S-Band tunable transformer has been built and measurements demonstrate a circuit which can transform impedances from 50 Ω to a range of 4 - 392Ω. A typical transformation from 50 Ω to 162 Ω yields a measured 0.5 dB instantaneous bandwidth of 14 - 16%, tunable bandwidth of 18%, and insertion loss of 0.6 - 0.9 dB. Applications for this novel technique include the design of electronically tunable wireless components, solid state power amplifiers, microstrip antennas, and microwave modulated laser transmit/receive systems.**

I. INTRODUCTION

THE problem of impedance matching is an important one which must be addressed in most microwave designs. Hybrid circuits are frequently matched using microstrip transmission lines whose dimensions are determined by considering a precise model of the load and a precise model of the matching network. Unfortunately, there are many times when accurate models and measurements are difficult to obtain because of nonlinear operation or measurement errors in very high Q circuits. In these cases, the microwave engineer must still resort to the "cut and try" approach to circuit design. Many electronically tunable phase shifters have been reported in the literature [1], but there is no mention of electronically tunable impedance transformers which are a required component in microwave matching. The following sections outline a technique for designing electronically tunable transmission lines capable of mimicking the performance of standard transmission lines but tunable in impedance and phase by imposing DC voltages. This is accomplished by utilizing GaAs varactor diodes separated by J-inverters in such a way as to simulate the performance of a transmission line over a specified bandwidth. It will be shown that the values of the required varactors and inverters can be simply related to the characteristic impedance and phase angle of a standard transmission line. A hybrid electronically tunable impedance transformer has been built using three varactor diodes separated by two quarter-wave microstrip lines. Through-reflect-line (TRL) calibrated measurements have been made on this circuit yielding surprisingly wide tuning range and bandwidth. The process of determining the precise voltages required to match a load is carried out automatically using a computer controlled closed loop feedback system utilizing the Simplex [2] optimization algorithm. Precise knowledge of parasitics in the electronically tunable impedance transformer are not required since they can be easily tuned out by closed loop computer control. Measurement results are presented and applications are discussed.

II. THEORY

It is well known that a transmission line can be approximated using the simple ladder network illustrated in Figure 1. The problem occurs when trying to tune this

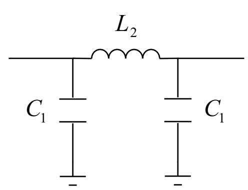

Figure 1. Lumped Element Transmission Line Approximation

network. Electronically tunable capacitors (varactors) are easily obtained, but electronically tunable inductors are not easily achieved except in a MMIC process [3-5]. Unfortunately, active tunable inductors have power

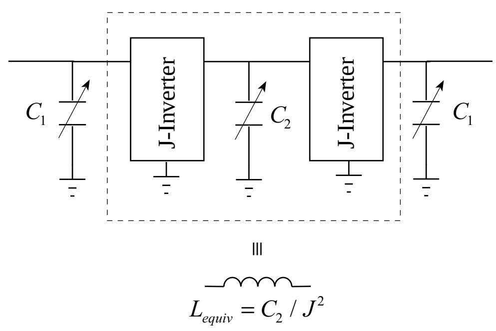

Figure 2. Electronically Tunable Line Topology

limitations which may be prohibitive in certain applications. While varactors are not high power devices, they can be used in medium power applications if biased appropriately [6]. We now consider the network illustrated in Figure 2. This circuit is electrically equivalent to Figure 1 over a specified bandwidth, and therefore can be used as an approximation for an electronically tunable transmission line. J-inverters are discussed in great detail by Matthaei *et.al.*[7]. To determine the required values for C_1, C_2 and J, the **Y** matrix for an ideal transmission line is equated to the **Y** matrix for the circuit in Figure 2. This is easily accomplished by first equating the ideal transmission line parameters to the parameters of the circuit in Figure 1. The steady state **Y** parameters for an ideal transmission line are as follows:

$$\mathbf{Y} = \frac{j}{Z_0}\begin{pmatrix} -\cot\frac{\omega\theta}{\omega_0} & \csc\frac{\omega\theta}{\omega_0} \\ \csc\frac{\omega\theta}{\omega_0} & -\cot\frac{\omega\theta}{\omega_0} \end{pmatrix} \qquad (1)$$

where ω is the frequency variable and ω_0 is the design frequency of the transmission line. Considering the **Y** parameters at the frequency of interest, we obtain

$$\mathbf{Y} = \frac{j}{Z_0}\begin{pmatrix} -\cot\theta & \csc\theta \\ \csc\theta & -\cot\theta \end{pmatrix} \qquad (2)$$

The steady state **Y** parameters for the circuit in Figure 1 are

$$\mathbf{Y} = j\begin{pmatrix} \omega C_1 - \frac{1}{\omega L_2} & \frac{1}{\omega L_2} \\ \frac{1}{\omega L_2} & \omega C_1 - \frac{1}{\omega L_2} \end{pmatrix} \qquad (3)$$

Equating terms in the **Y** matrices of (2) and (3) the two following unique equations are obtained:

$$C_1\omega - \frac{1}{L_2\omega} = -\frac{\cot\theta}{Z_0} \qquad (4)$$

$$\frac{1}{L_2\omega} = \frac{\csc\theta}{Z_0} \qquad (5)$$

Assigning the variables α and β to the right sides of (4) and (5) respectively, it is simple to solve for C_1 and L_2 as follows:

$$C_1 = (\alpha + \beta)/\omega \qquad (8)$$

$$L_2 = \frac{1}{\omega\beta} \qquad (9)$$

Finally, using the transformation $L_2 \rightarrow C_2 / J^2$ as illustrated in Figure 2, the expressions for C_1 and C_2 are obtained as follows:

$$C_1 = \frac{1 - \cos\theta}{Z_0\omega\sin\theta} \qquad (10)$$

$$C_2 = \frac{J^2 Z_0 \sin\theta}{\omega} \qquad (11)$$

Conversely, one can solve for Z_0 and θ in terms of C_1 and C_2 as follows

$$\theta = \arccos\left(1 - C_1 C_2 \omega^2 / J^2\right) \qquad (12)$$

$$Z_0 = \frac{C_2\omega / J^2}{\sqrt{1 - \left(1 - C_1 C_2 \omega^2 / J^2\right)}} \qquad (13)$$

It is interesting to note that for (12) and (13) to yield a valid solution,

$$\frac{J^2}{\omega^2} > C_1 C_2 \qquad (14)$$

It is clear that not all possible values of C_1 and C_2 yield a circuit equivalent to a transmission line. Equations (10) and (11) in conjunction with (14) provide a powerful yet simple way to compute the required components to synthesize a tunable impedance transformer over any desired range of frequencies and transformation ratios.

III. DESIGN EXAMPLE AND MEASURED RESULTS

As an illustration of this new technique of designing electronically tunable microwave impedance transformers, a hybrid circuit was built using three varactor diodes in conjunction with two 50 Ω microstrip quarter wave transformers. The quarter wave transformers serve as the required J-inverters discussed in section II. This circuit was fabricated using Rogers thermoset plastic (TMM10i™). The varactors were MA46473 GaAs hyperabrupt junction (γ=1.25) diodes manufactured by MA-COM. These diodes have a quality factor of 3000 at -4 volts measured at 50 MHz. The circuit layout is

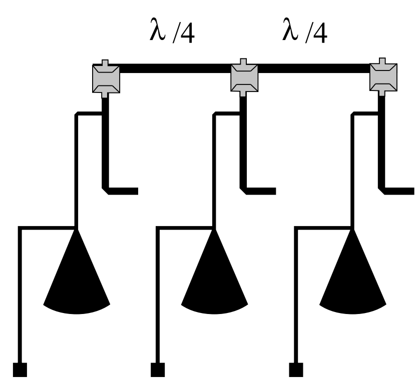

Figure 3. S-Band Electronically Tunable Impedance Transformer

illustrated in Figure 3. The circuit was designed using completely distributed bias tees to decrease system losses and to avoid the need for plated through holes. The goal in this approach was to provide a low resistance DC bias suitable for medium power solid state amplifier input matching (15 - 25 dBm). The network biasing has been designed so that the gate bias for a GaAs FET can be applied directly to this network. It is important to note that the size of this network can be greatly reduced by replacing the λ/4 wavelength lines with reduces length lines and fixed capacitors as described by Hirota, *et. al.* [8]. Using this technique the size of this tunable impedance transformer can be reduced by as much as 80% as claimed in [8].

The best approach for measuring the performance of the impedance transformer is to look at the insertion loss in a mixed impedance system. After all, the purpose of the transfomer is to transfer as much power from the source at one impedance to the load at a second impedance. This insertion loss can be obtained by observing the value of S_{21} in a mixed impedance system (i.e. input impedance = 50 Ω and output impedance = 8 Ω). These mixed impedance S-Parameters can be derived directly from standard 50 Ω S-Parameters by carrying out a simple renormalization.

If the system is well matched, and therefore S_{11} and S_{22} are very small, then S_{21} will be determined predominately by the resistive loss in the system, which is to be kept at a minimum. Figure 4 illustrates the measured performance of the electronically tunable microwave impedance transformer shown in Figure 3. Each curve is obtained by renormalizing the measured S-Parameters to the specified source and load impedances in the legend. These curves were obtained by allowing a computer to repeatedly read the measured circuit S-parameters, compute an error function, and calculate new bias voltages for the circuit thereby iteratively attempting to minimize

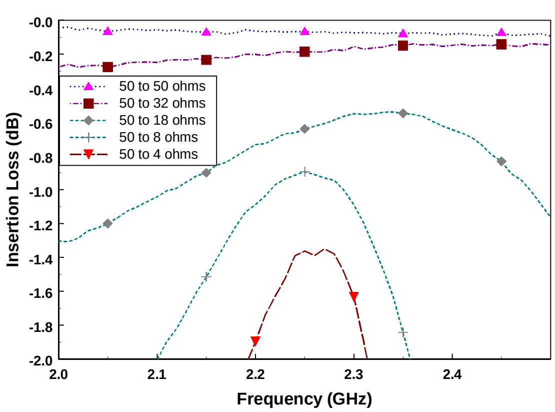

Figure 4a. Measured Insertion Loss of an Electronically Tunable Microwave Impedance Transformer - Low Impedances

the insertion loss in the specified mixed impedance system. This automated control was accomplished by using MMICAD™ for Windows by Optotek in conjunction with an HP8510C vector network analyzer and two GPIB controllable Hewlett Packard power supplies. The curves in Figure 4 illustrate the minimum insertion loss values obtainable at 2.25 GHz, the center frequency of this design. The curves are illustrated over frequency to illustrate the instantaneous bandwidth characteristics of the system which is in excess of 4.8%. Further measurements were made above and below the center frequency to obtain the tuning bandwidth of the system. Figures 5 a, b, and c illustrate measured performance of the electronically

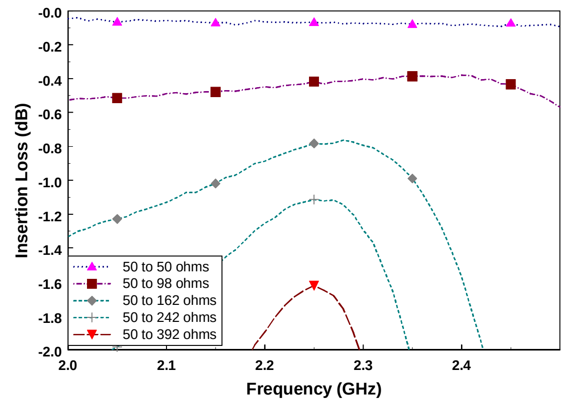

Figure 4b. Measured Insertion Loss of an Electronically Tunable Microwave Impedance Transformer - High Impedances

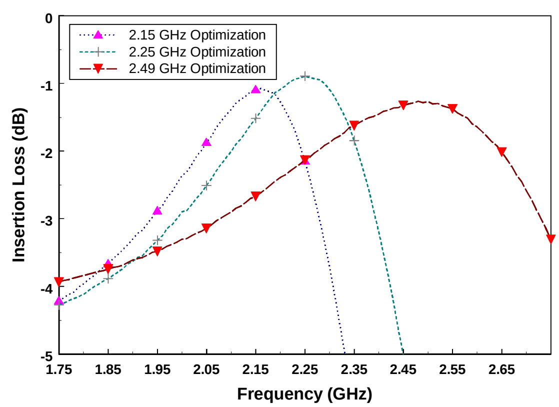

Figure 5a. Measured 50 Ω to 8 Ω Impedance Transformation Optimized at 2.15, 2.25 and 2.49 GHz

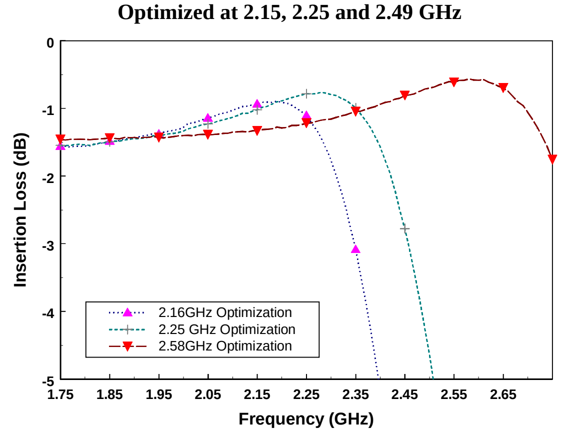

Figure 5b. Measured 50 Ω to 162 Ω Impedance Transformation Optimized at 2.16, 2.25 and 2.58 GHz

tunable transformer after optimizing performance at frequencies as far apart as possible above and below the design frequency of 2.25 GHz. This tunable bandwidth is in excess of 10% for all illustrated transformations. Figures 6 and 7 illustrate the performance of the electronically tunable transformer by illustrating the minimum obtainable insertion loss when matching a range of real impedances to 50 ohms. It is important to note that the optimization was done using minimum insertion loss as the criterion for determining the error function. This

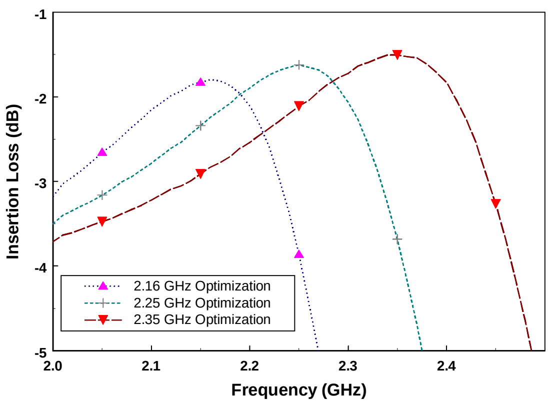

Figure 5c. Measured 50 Ω to 392 Ω Impedance Transformation Optimized at 2.16, 2.25 and 2.35 GHz

explains the somewhat unusual behavior that is seen in the range of varactor voltages and insertion phase in Figure 7. The computer was able to trade off the ohmic losses in the system with reflected loss in such a way as to obtain minimum insertion loss. Part of the beauty of this electronic feedback control is that this otherwise analytically complex task becomes relatively simple to implement.

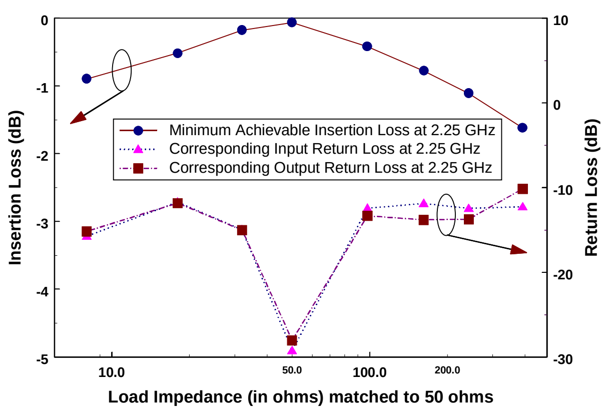

Figure 6. Measured Minimum Insertion Loss and Corresponding Return Loss of the Computer Controlled Electronically Tunable Microwave Impedance Transformer at 2.25 GHz

IV. CONCLUSION

A new method for designing an electronically tunable microwave impedance transformer has been demonstrated. A 2.25 GHz hybrid electronically tunable transformer has been fabricated and tested using microstrip technology. Results show a transformation range of 4 Ω to 392 Ω from a 50 Ω source. A minimum instantaneous bandwidth of 4.8% and a minimum tunable bandwidth of 10% has been demonstrated over this entire range of transformations.

Due to the demonstrated low insertion loss and wide tunable bandwidth, this new type of electronically tunable impedance transformer has many applications including computer controlled and adaptive matching of medium power solid state amplifier inputs, wireless components, photodetectors, and laser diodes. Additionally, this new matching circuit building block can be used in conjunction with an electronically tunable phase shifter to provide impedance matching of poorly characterized complex loads. The use of this electronically tunable transformer should bring the microwave community one step closer to eliminating the common practice of cutting microstrip with a knife to impedance match certain difficult to model microwave circuits.

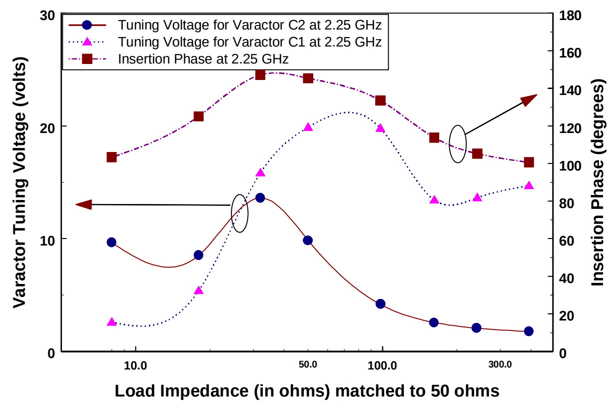

Figure 7. Measured Varactor Voltages and Insertion Phase of the Computer Controlled Electronically Tunable Microwave Impedance Transformer at 2.25 GHz

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