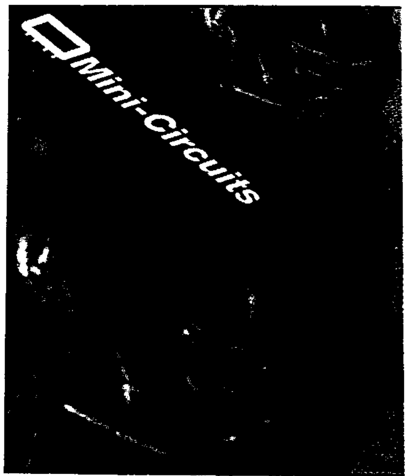
microwave JOURNAL®

JUNE 2001 • VOL. 44, NO. 6



ON THE COVER

A new series of low cost, high quality directional couplers is featured on this month's cover

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Microwave Journal (USPS 396-250) (ISSN 0192-6225) is published monthly by Horizon House Publications Inc. 685 Canton St., Norwood, MA 02062. Periodicals postage paid at Norwood, MA 02062 and additional mailing offices.

POSTMASTER: Send address corrections to *Microwave Journal*, PO Box 3256, Northbrook, IL 60065-3256. Subscription information: (847) 291-5216. This journal is issued without charge upon written request to qualified persons working in that part of the electronics industry, including governmental and university installation, that deals with VHF through light frequencies. Other subscriptions are: domestic, $115.00 per year, two-year subscriptions, $175.00; foreign, $190.00 per year, two-year subscriptions, $355.00; back issues (if available) and single copies, $10.00 domestic and $20.00 foreign.

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TECHNICAL FEATURE

HIGH SPEED ANALYSIS AND OPTIMIZATION OF WAVEGUIDE BANDPASS FILTER STRUCTURES USING SIMPLE NEURAL ARCHITECTURES

At microwave frequencies, from about 7 to 60 GHz, inductive irises are very often used as coupling networks between half-wavelength cavities in rectangular waveguides to develop very selective low loss bandpass filters. This is due to the fact that symmetric and asymmetric metal inserts, along with small tuning posts, are very easy to manufacture in large production volume. To facilitate the design and optimization process, a simple and very accurate neural architecture is presented, which is easily translated to a standard electrical equivalent circuit that reproduces in a wide range of iris aperture, thickness and frequency. The proposed new models, although they can be embedded into any commercial microwave software, have been easily implemented into MMICAD® Comparisons have been made for high order high frequency waveguide half-wave filters, showing an excellent agreement with full three-dimensional (3D) electromagnetic HP-HFSS® simulations along with computation speeds thousands of times faster.

To the authors' knowledge, none of the current commercial microwave aided design (CAD) programs incorporate models for useful discontinuities in rectangular waveguide, while they have models for the standard TE10 waveguide transmission line. This means that, for a microwave designer, the waveguide world belongs to other kinds of simulators based on hard numerical methods — mode matching and finite elements where the design and optimization cycles are still very long.

In the case of planar structures (microstrip, strip and coplanar lines), the microwave designer has a wide range of electrical models available (sometimes based on electromagnetic simulations) for almost any discontinuity in such transmission media, and the designer can verify or fine-tune his or her final design through the use of accurate 2D and 2.5D planar electromagnetic simulators. The same concept can be extended to the waveguide world for some useful

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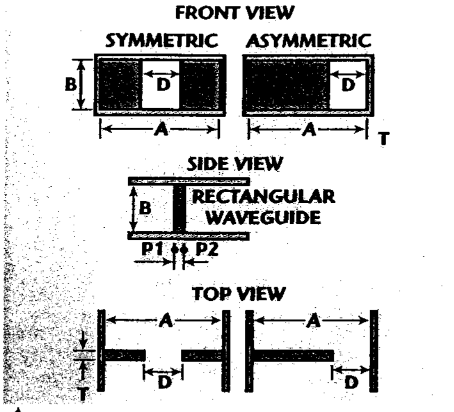
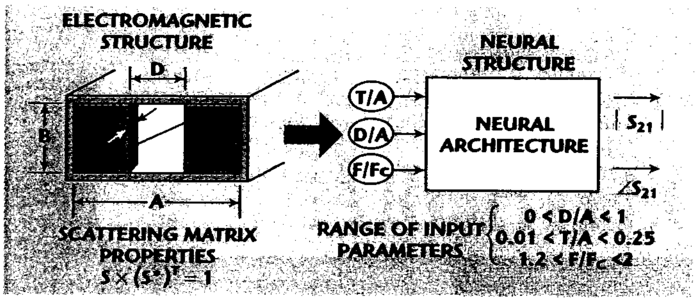
well-known discontinuities such as symmetric and asymmetric irises for microwave bandpass filter design shown in ***Figure 1***.



▲ *Fig. 1 Symmetrical and asymmetrical rectangular waveguide inductive iris.*



▲ *Fig. 2 Proposed equivalence between electromagnetic and neural structures.*

The existing circuit models for waveguide inductive irises, starting from Marcuwitz's work,[1] are published elsewhere; however, most of them are developed in terms of recursive closed form equations coming from electromagnetic pseudo quasi-static and full-wave approaches. Due to the multimode dispersive nature of these electromagnetic discontinuities, the equations are not easily implemented into commercially available circuit simulators. Furthermore, they are rather tedious and computation intensive, thus preventing an easy filter analysis and optimization process. Their frequency accuracy for a single iris is perhaps sufficient, but when using a high order filter structure, the propagation of the individual errors through the filter gives poor results (for example, bandwidth shift and in band attenuation). The second available solution, the use of full 3D electromagnetic simulators such as HP-HFSS®,[2] is accurate, but unacceptable in computation time (several hours/days) when used for filter design and optimization.

Instead of searching for more precise, and therefore more complicated closed form equations, the idea proposed here is to use simple and accurate neural architectures to fit the scattering parameters obtained by using a precise full 3D simulator for single and double inductive irises in rectangular waveguides. Since the electromagnetic discontinuity of a single or double inductive iris of aperture D and thickness T in a TE10 propagating rectangular waveguide behaves like a lossless symmetrical two port reciprocal network at the reference planes P1 and P2, it is enough to adjust a single two port parameter at the output of the neural network. For microwave filter applications, it is convenient to control the forward scattering parameter S_{21} (easily related to the traditional Z or Y parameters), and furthermore to use the well-known properties of the scattering matrix to derive the other Sij parameters as

$$S \times (S^*)^T = I \qquad (1)$$

At this point, it is evident that the input parameters for a possible neural structure should be the physical dimensions of the inductive iris, that is, D and T, along with the waveguide dimensions (A and B) and the frequency of operation. This primary strategy exhibits some important disadvantages because the standard waveguide dimensions are defined for precise frequency bands where the TE10 is the dominant propagating mode, and in a first approach a neural topology should be derived for each waveguide band (which is not a general approach). However, if the scaling properties of the waveguide structures are considered, the normalised iris dimensions D/A and T/A can be used as input parameters, as well as the normalised frequency F/Fc, where Fc is the cut-off frequency of the TE10 mode, as shown in ***Figure 2***.

Furthermore, the range of the input parameters should have some constraints regarding the usual wave-

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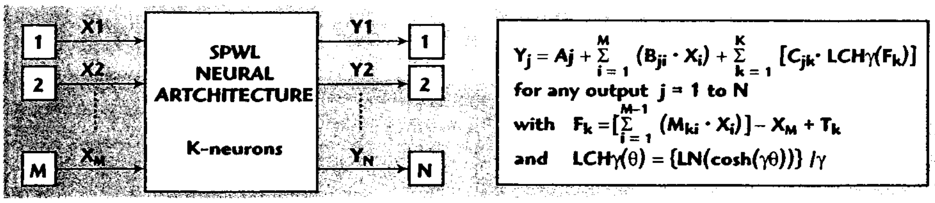

▲ *Fig. 3 General SPWL neural architecture.*

guide filter utilisation. The iris aperture could vary from 0 to the maximum aperture A, that is, $0<D/A<1$, while a reasonable range for the iris thickness T should be given by $0.01<T/A<0.25$. Finally, the normalised frequency band should be $1.2<F/F_c<2$ in order to avoid unwanted propagation modes. In conclusion, this general strategy uses a single neural architecture for the S_{21} parameter, and three normalised parameters as input data. It should be enough for any given frequency band where this kind of filter is applicable.

THE NEURAL ARCHITECTURE

From an intuitive point of view, a neural network[3] can be viewed as a parallel distributed processor that exhibits a natural ability for storing experimental knowledge and making it available for ulterior use. This knowledge is acquired through dedicated learning algorithms, along with a weighty interneuron connection. The typical neural networks, MLP and RBF families, normally require a relatively large number of neurons for a close fit to the experimental data. Because the objective is to be highly competitive against the pure electromagnetic simulation, a new SPWL[4] has been chosen for this particular problem.

The proposed SPWL model is an extension of the well-known canonical piecewise linear model (PWL) described by Chua.[5] In its basic formulation, the Canonical PWL model performs any general nonlinear mapping $F : R^M \rightarrow R^N$ (M inputs and N outputs) by means of the expression

$$Y = F(X) = A + B \bullet X + \sum_{k=1}^{K} C_k \left| < \alpha_k, X > - \beta_k \right| \quad \text{(for K neurons)} \qquad (2)$$

where

$X(M)$ = input vector
$Y(N)$ = output vector
$A(N)$, $B(N \times M)$, $\alpha_k(M)$, $C_k(N)$ = fitting vectors
β_k = scalar
$<\alpha_k, X>$ = inner product

This model divides the input space into different regions by means of several boundaries implemented by hyperplanes of dimension M-1. It then constructs the function approximation by means of a combination of hinging hyperplanes of dimension M. Such hinging hyperplanes are the result of joining two linear hyperplanes over the boundaries defined in the input space.

It can be seen that the expression inside the absolute value function defines the boundaries partitioning the domain space. This function controls the transition between linear regimes and, therefore, the Canonical PWL model inherits some properties from the absolute value function; it is continuous but not derivable along the boundaries. Moreover, the second and higher order derivatives are zero except at the boundaries where they are discontinuous, which is critical for circuit optimization purposes. To overcome this drawback, the substitution of the absolute value function is proposed for a derivable function in order to smooth the joint of hyperplanes at the input space boundaries. Several possibilities exist to smooth the absolute value function allowing, at the same time, a parametric control of the "sharpness" of the transition. The smoothing function is chosen as

$$LCH\gamma(\tau) = \frac{Ln\left[\cosh(\gamma\tau)\right]}{\gamma} \qquad (3)$$

where

γ = parameter that allows the smoothness of the transition to be controlled

The advantage is clear when one looks for the derivative of Equation 3: $d/d\tau\{LCH\gamma(\tau)\} = \tanh(\gamma\tau)$ which is the activation function of a universal approximator such as the MLP. ***Figure 3***

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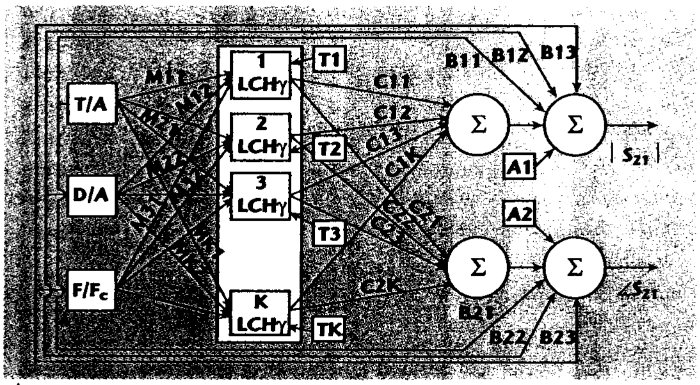
shows a descriptive view of the proposed SPWL model.

MODEL VALIDATION

The above description has been applied to the electromagnetic structures, that is, both symmetric and asymmetric irises. This model provides a smooth and derivable approximation that improves considerably the performance of the Canonical PWL model when it is applied to real microwave devices, mainly in the optimization process. Moreover, it requires a much smaller number of parameters and a lower computation burden than other models commonly used. Extensive full 3D electromagnetic simulations have shown that the proposed architecture, shown in ***Figure 4***, is able to reproduce the two port complex S_{21} parameter for a wide range of input data - (0.01 < T/A < 0.25), (0 < D/A < 1) and (1.2 < F/F_c < 2.0), thus covering most applications. In this case, a very good individual iris fit by using a seven-order SPWL is obtained; the maximum error in S_{21} for any individual iris, when compared with HFSS simulation, is less than 0.02 in module and less than 2° in phase. ***Figure 5*** shows as an example the neural fitting parameters for the symmetric iris case along with a comparison between full 3D electromagnetic simulation and the neural model for a relative large iris thickness (T/A = 0.14) and for various iris apertures D/A as a function of the normalised frequency.

Although it is very easy to show how the proposed method accurately fits the frequency behaviour of an individual inductive iris, when designing high order microwave filters, the propagation of the individual errors could be important, especially for very narrow bandpass filters. This fact is a higher level test of the validity of the approximation. For

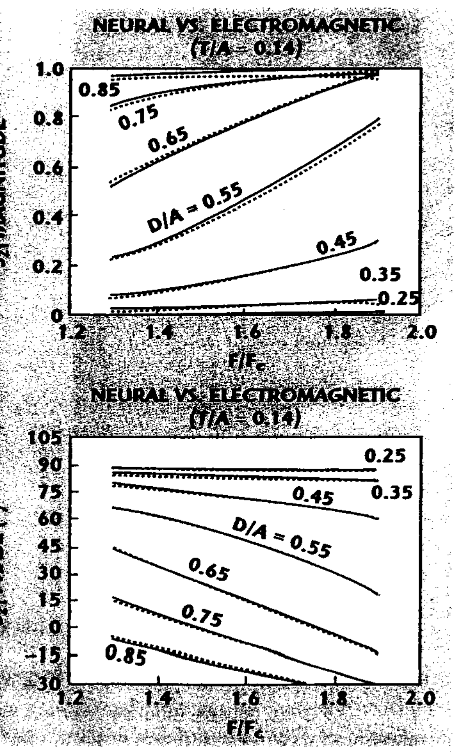


▲ *Fig. 4 Dedicated neural architecture for symmetric and asymmetric waveguide irises.*

Fig. 5 Single symmetrical iris validation. ▼

7-ORDER NEURAL PARAMETERS FOR SYMMETRIC IRIS

A1 = + 0.5313133	A2 = + 72.029501	γ = + 6.5000
B11 = − 0.0224248	B21 = − 6.3434002	M11 = − 0.1471183
B12 = − 0.0307383	B22 = − 118.99504	M12 = + 1.1227934
B13 = − 0.0096260	B23 = − 33.658561	M21 = + 0.3671070
C11 = + 2.0500217	C21 = − 64.315213	M22 = + 4.1136844
C12 = − 0.0034359	C22 = − 18.128299	M31 = − 0.1056092
C13 = − 4.8298888	C23 = + 289.59198	M32 = + 0.2939928
C14 = − 0.1482995	C24 = + 57.938470	M41 = + 0.2864985
C15 = − 1.5116727	C25 = + 92.329060	M42 = + 0.0385768
C16 = + 4.3496077	C26 = − 403.68554	M51 = − 0.3388533
C17 = + 0.1129634	C27 = + 47.962584	M52 = + 0.7186874
		M61 = + 0.1528574
		M62 = + 0.0087490
		M71 = − 0.2991391
		M72 = + 2.2288361
		T1 = + 0.4822774
		T2 = − 0.2073612
		T3 = + 0.6221584
		T4 = + 0.0607542
		T5 = + 1.0709634
		T6 = + 0.7017891
		T7 = + 1.0711678



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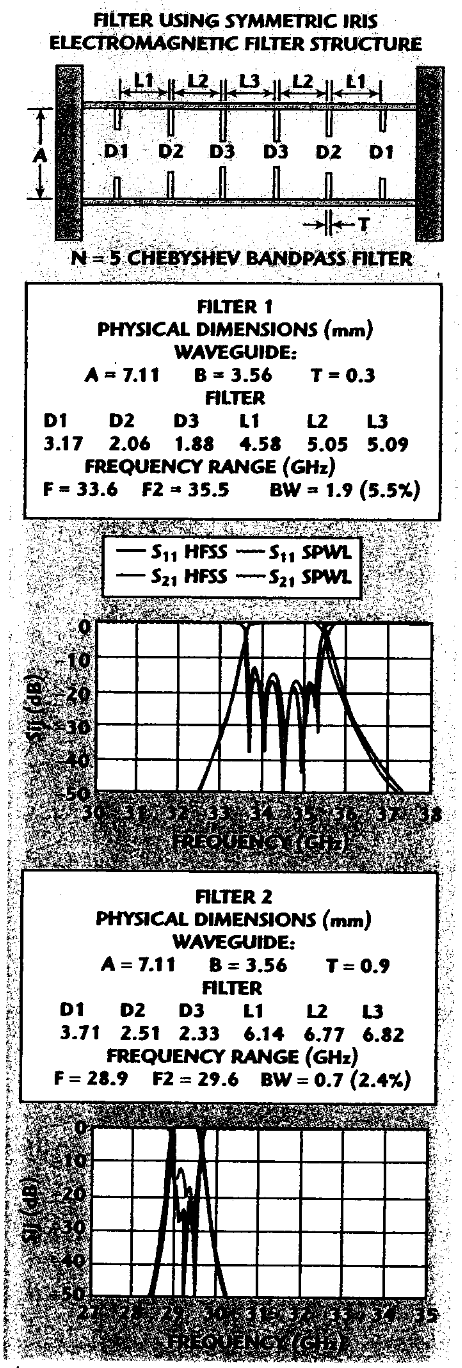
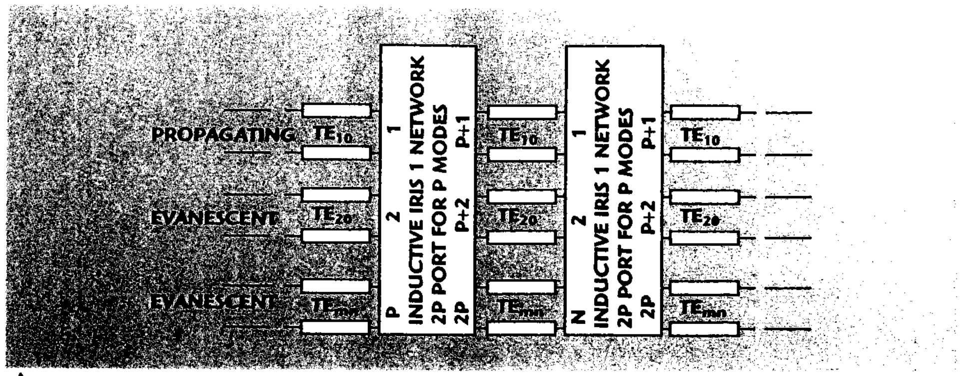
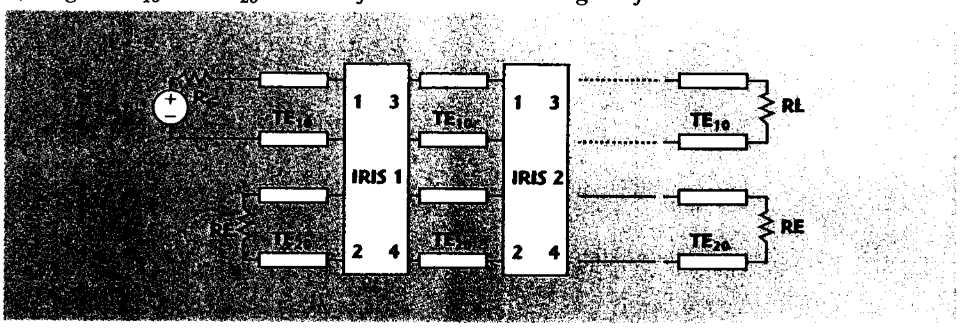


▲ *Fig. 6 Model validation through filter implementation.*



▲ *Fig. 7 Multimode structure for an iris-based waveguide filter.*

▼ *Fig. 8 TE_{10} and TE_{20} structure for an iris-based waveguide filter.*



this reason the neural architecture has been implemented easily into MMI-CAD[6] by using its MDL capability along with the flexibility in working with electrical model and local variables. The individual irises are joined by using fundamental TE10 waveguides (available in any simulator). At least for symmetric iris structures, and for these half-wave filters, it is not necessary to take into account high order connecting modes. Up to 21 different multi-section Chebyshev/Butterworth bandpass filters in different waveguide bands were tested, always showing very good agreement with full 3D electromagnetic simulations and having a computing simulation time more than 1000 times faster than any conventional analysis. Furthermore, the filter optimization process takes only a few seconds. This is due to the fact that the chosen algorithm is not only very fast but also continuous in its high order derivatives.

Figure 6 shows the general structure for microwave half-wave filters that use double inductive irises in a waveguide environment. For validation purposes, WR22 Ka-band waveguide (26.5 to 40.0 GHz) is chosen, where two very different N = 5 (6 iris discontinuities) Chebyshev bandpass filters have been designed and optimized. Filter 1 uses symmetrical irises having moderate (0.3 mm) thickness, with a center frequency at f_0 = 34.55 GHz and having a fractional bandwidth of 5.5 percent. Conversely, Filter 2 is a very narrow band waveguide filter (2.4 percent fractional bandwidth) centered at f_0 = 29.25 GHz that uses very thick (0.9 mm) irises. For both cases, all the physical dimensions are shown in the figure. In terms of analysis, HP-HFSS means full 3D electromagnetic simulation and SPWL means neural electrical equivalent circuit simulation. Note that the model implementation is extremely robust, even for very narrow filters, and it is difficult to distinguish between the two simulations.

HIGH ORDER TE10 + TE20 IRIS CONNECTION

From a general point of view, a waveguide having P modes should be considered the connecting media between successive discontinuities, that is, the discontinuity should be described as an electrical 2P-port characterised by its generalized multiport scattering matrix S(2Px2P), as shown in ***Figure 7***. Since the structure of a symmetric iris exhibits perfect symmetry, only odd modes can be excited at the discontinuity, that is, the first high order mode to be considered is the TE30. Multimode electromagnetic simulations show that for half-wave waveguide filters that use symmetric iris structures, it is enough to consider only the first connecting mode TE10, as can be seen from the results obtained for the filters using symmetric irises.

Unfortunately, this is not the case for half-wave waveguide filters that use asymmetric iris structures where the above approach is not accurate enough. Due to the non-symmetrical nature of these discontinuities, a non-negligible contribution of the TE20 mode along with an insignificant contribution of the remaining high order connecting modes can be expected. Extensive filter simulations corroborate this assertion and the final filter structure is shown in ***Figure 8***.

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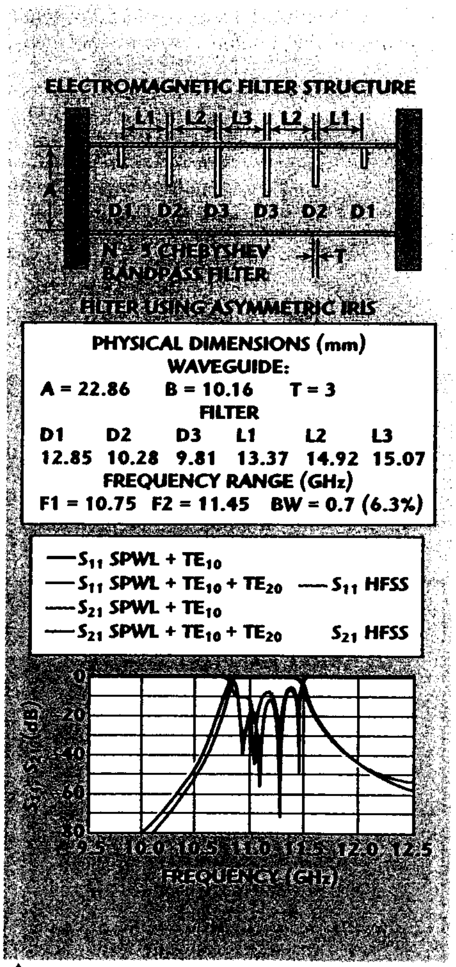
At this point it should be kept in mind that the generalized scattering matrix for the first two modes of a single iris is a 4x4 matrix having some special properties.[7] The $S_{11} = S_{33}$ and $S_{13} = S_{31}$ elements belong to the propagating mode TE10; they have the matrix properties shown in Equation 1. The terms $S_{14} = S_{41} = S_{23} = S_{32}$ and $S_{12} = S_{21} = S_{34} = S_{43}$ relates the propagating TE10 mode with the evanescent TE20 mode. After manipulation of the scattering properties they can be related in a simple manner, as shown in Equation 4.

$$S_{14} = \frac{-j\dot{S}_{12} - \dot{S}_{11}S_{12}}{\dot{S}_{13}}$$

or

$$S_{12} = \frac{-j\dot{S}_{14} - \dot{S}_{11}S_{14}}{\dot{S}_{13}} \quad (4)$$

Finally, mode matching simulations show that the terms $S_{22} = S_{44}$ and $S_{24} = S_{42}$ can be made zero without any loss of accuracy because these elements relate to evanescent modes only.



▲ *Fig. 9 HFSS vs. SPWL-TE$_{10}$ and SPWL-TE$_{10}$ + TE$_{20}$ comparison.*

In conclusion, the only need is to develop an extension of the initial neural network that has as outputs the magnitude and phase of S_{31}, along with S_{14}, for example. The remaining elements of the 4×4 scattering matrix can be deduced from the above equations. This SPWL architecture is easily converted into a standard electrical equivalent circuit and implemented into any circuit simulator and then used in the same manner as the element "bend" or "step" in microstrip, for example. The final point is to build an electrical model for the evanescent TE20 waveguide mode; however, this is not a problem because its Z matrix is well-known in the literature.

As a validation example of the aforementioned theory, ***Figure 9*** shows a comparison for an X-band (6 percent BW) half-wave filter that uses asymmetric iris structures. As shown, it is almost impossible to distinguish between the multimode HFSS simulation and the proposed 4-port SPWL approach. The differences are in the range of the mechanical tolerances. However, there is a 1 percent frequency shift when using the simple 2-port SPWL approach, which is unacceptable for a 6 percent filter bandwidth, thus confirming the dual mode assumption.

CONCLUSION

A very simple and extremely accurate SPWL neural architecture for inductive iris in electromagnetic structures has been presented. In order to cover the whole microwave range, normalized physical dimensions and frequency have been used as input parameters to the network. Model implementation has been accomplished through the use of standard electrical equivalent circuit structures. Model validation has been achieved for very different high order microwave filter applications, always showing excellent agreement when compared with the well-known accuracy of a full 3D electromagnetic simulator. Since the neural architecture is continuous in its high order derivatives, the filter optimization process can easily be accomplished. Finally, simulations have shown that the proposed strategy is more than 1000 times faster than any commercially available electromagnetic simulator, thus allowing the microwave engineer to really minimize the microwave filter design process without loss of accuracy. ■

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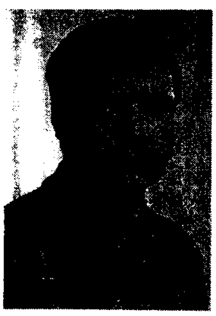

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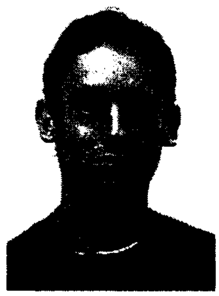

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