At least $N + 1$ Finite Transmission Zeros Using Frequency-Variant Negative Source-Load Coupling

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***Abstract*—A second-order cross-coupled combline filter which has three finite transmission zeros is presented. The problem of the frequency-invariant coupling in a real circuit is introduced. To make extra transmission zeros, a top metalized dielectric block is used.**

***Index Terms*—Bandpass filters, combline filters, cross coupling, transmission zeros.**

I. INTRODUCTION

COUPLING matrices of the low-pass prototype filter have been widely used in direct-coupled resonator bandpass filter synthesis. Frequency-invariant coupling matrices for the given topology can be obtained by using a low-pass prototype filter and ideal impedance inverters, combined with a sequence of similarity transformations and efficient optimization techniques [1]–[4]. In implementing the coupling matrix, filter response is degraded to some extent and tuning and optimization are required. Sometimes coupling coefficients from the coupling matrix may not be realizable. Under the frequency-invariant coupling, $n - 2$ is the maximum number of finite transmission zeros. This is theoretically verified for n-coupled resonator networks without source-load coupling [5]. In recent work [6], when source-load coupling is involved, realization of at least n maximum finite transmission zeros from n-coupled resonators is presented.

However, the result is only valid for frequency-invariant couplings. With a frequency-variant inverter, a waveguide cavity filter that has more than n transmission zeros was reported [7]. In this letter, the difference between the frequency-variant coupling and invariant coupling in filter response is introduced. An available method for source-load negative coupling in planar bandpass filter design is proposed. This result shows three finite transmission zeros out of two physical resonators. A two-pole combline filter with equivalent circuit is presented as an example.

II. COMPARISON OF THE FREQUENCY–INVARIANT AND FREQUENCY–VARIANT COUPLINGS

The approach using the frequency-invariant coupling matrix is quite accurate for the narrow band filter synthesis. However, this method does not show filter response in the rejection band in an accurate manner [8]. A two-pole bandpass filter, which has a negative cross coupling from source to load, is shown in

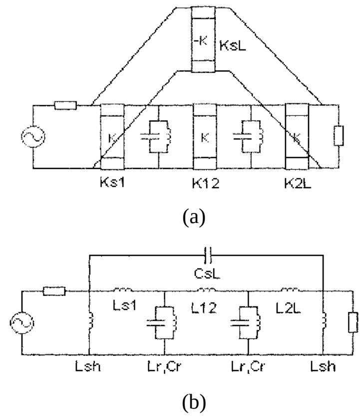

Fig. 1. Negative source-load coupling two-pole filter. (a) Frequency-invariant ideal inverter negative source-load coupling ($-K$); (b) negative frequency-variant source-load coupling after Pi-inverter transformations.

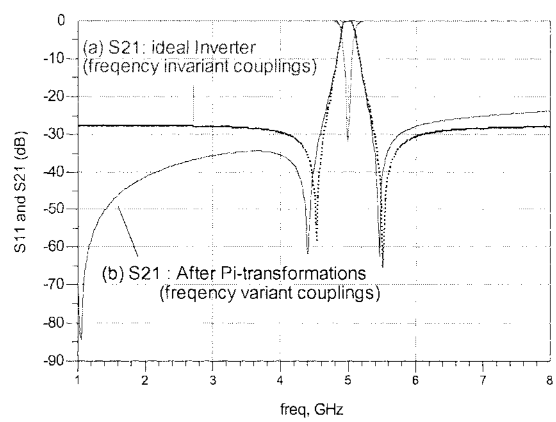

Fig. 2. Comparison of source-load cross-coupled two-pole filter with a negative source-load coupling between frequency-invariant and frequency-variant couplings.

Fig. 1. Impedance inverters are used to show the frequency-invariant couplings and to make a negative source-load coupling ($-K$) with respect to the main coupling. An achievable practical circuit after Pi-transformation of ideal inverters is shown in Fig. 1(b). When a positive source-load coupling is employed, two finite transmission zeros (FTZs) can be obtained, i.e., n maximum FTZs out of n resonators. The negative coupling interacts with the main coupling elements and results in $n + 1$ FTZs (Fig. 2).

As seen in Fig. 2, near the center frequency of the filter, both cases are well matched. However, the filter response is quite different in the stopband frequencies. Thus, the truly frequency-variant couplings must be taken into account to correctly predict the filter stopband performance. We can take advantage of this

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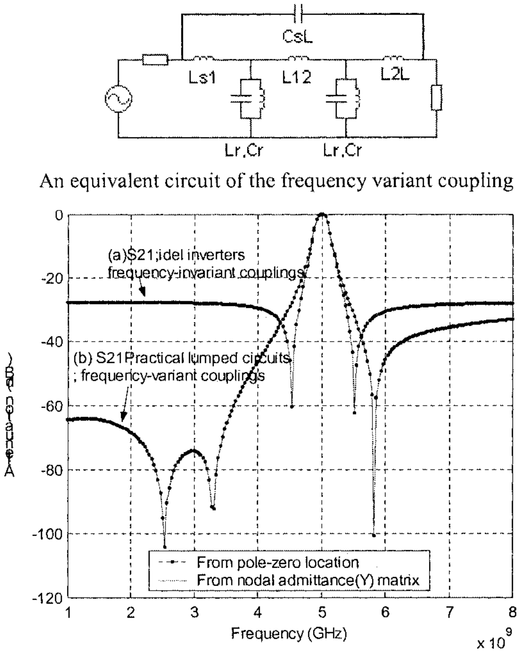

Fig. 3. Filter response of the frequency-invariant and frequency-variant couplings.

phenomenon to improve filter performance, by moving the extra TZ closer to the passband.

The negative cross coupling (capacitive) and main coupling (inductive) make at least three FTZs on the imaginary axis and show asymmetrical insertion loss in the rejection band (Fig. 3). To get closer, lower side two FTZs, shunt inductors (L_{sh}) from the input/output transformer sections are removed and the original circuit of Fig. 1(b) is tuned to obtain the response as shown in Fig. 3. Since coupling coefficients are actually frequency-variant, evaluation over the wide frequency range shows that ideal frequency-invariant inverters do not correctly predict extra FTZs in the rejection band.

For any filter network, the pole and zero locations can be calculated from the network function. Frequency response of a network is the value of the network function $H(s)$ at $s = j\omega$

$$|H(s)| = \frac{1}{c} \frac{\prod_{i=1}^{m} |(s - z_i)|}{\prod_{i=1}^{n} |(s - p_i)|} \tag{1}$$

where c is a scaling factor, and p_i, z_i are pole and zero locations.

Fig. 3 compares the filter response of the frequency-invariant and frequency-variant networks from pole and zero locations and the nodal admittance (Y) matrix [9]. Three FTZs explicitly show the effect of the frequency-variant coupling contrasted to the frequency-invariant case. In reality, the finite Q value of each element only affects the real part of pole and zero locations, not the number of poles and zeros. It is expressed as the degradation of insertion and return loss (Fig. 4). The pole and zero locations for finite Q networks clearly shows the existence of the extra transmission in the lower stop band (Fig. 5).

For reasonably high Q values, it is easy to apply the frequency-variant coupling approach to a practical circuit, controlling the TZs location in the stopband. Each FTZ is correlated

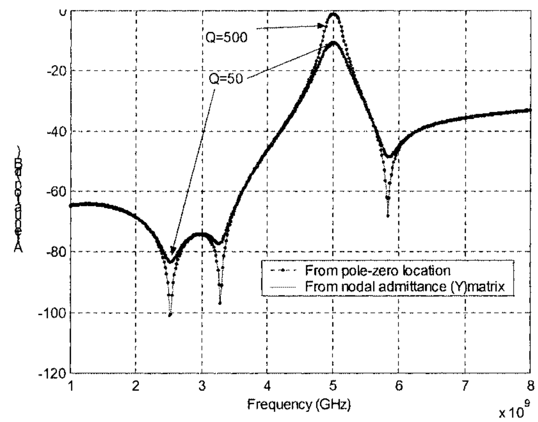

Fig. 4. Insertion and return loss with three FTZs after tuning.

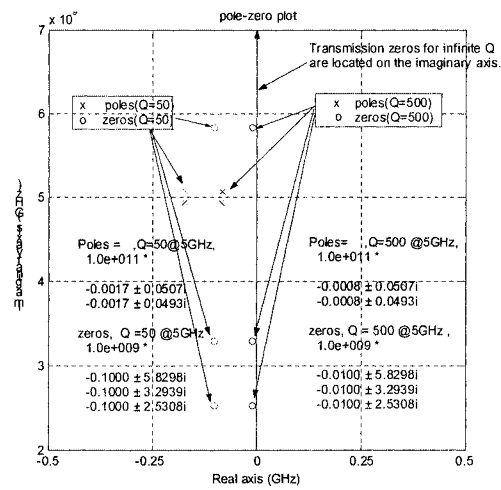

Fig. 5. Pole and zero locations of the frequency-variant couplings in Fig. 3 circuit. Poles and zeros on the left are for $Q = 50$; poles and zeros on the right are for $Q = 500$.

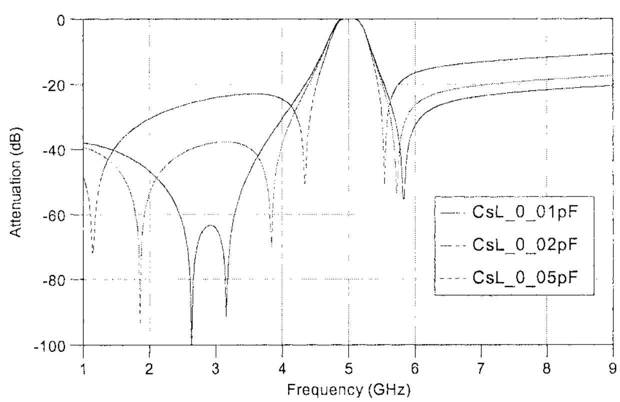

Fig. 6. Effect of the source-load bridging capacitance (C_{SL}) on the transmission zeros in Fig. 3 circuit.

with the bridging capacitor (C_{SL}) and series inductors (L_{S1}, L_{12}, L_{2L}) in Fig. 3 circuit. Thus, these FTZs are not controllable separately in this lower-order configuration. FTZs are controllable in a limited range, while maintaining pass band characteristics. Fig. 6 shows the effect and limitation of the bridging capacitance on the transmission zeros.

In the following example a combline filter is used (Fig. 7). A second-order example filter is shown to achieve three finite

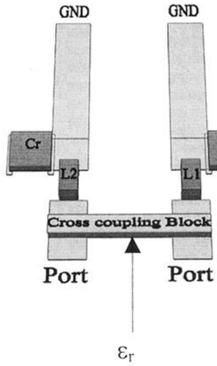

Fig. 7. Two-pole combline filter, three finite TZs, using the top metalized dielectric block for source-load coupling.

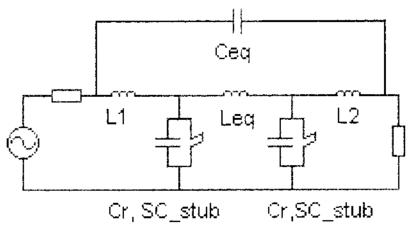

Fig. 8. Equivalent circuit of Fig. 7, where $Ceq = (\varepsilon_o \varepsilon_r * A/t)$; $Leq =$ net coupling between parallel-coupled lines; A: cross-sectional top and bottom metal size; t: dielectric substrate thickness.

TABLE I
VALUES OF COMPONENTS FOR CIRCUITS OF FIGS. 1, 3, AND 7 ($Rs = R_L = 50\ \Omega$), INVERTERS ARE CONVERTED AT $Fc = 5$ GHz

Ideal inverters(Fig.1.a)	Real Circuit (Fig. 1.b)	Tuned Circuit (Fig. 3)	Cross-coupled Combline Bandpass filter (Fig. 7 & Fig. 8)
$K_{S1} = K_{2L} = 237.21$	$L_{S1} = L_{2L} = 7.381$ nH	$L_{S1} = L_{2L} = 1.7799$ nH	$L_1 = L_2 = 8.4$ nH, Q=1200 @ 5GHz
$K_{12} = 1125.3$	$L_{12} = 35.82$ nH	$L_{12} = 3.582$ nH	Line gap = 90 mils, (Leq=44 nH)
$K_{SL} = -2400$	$C_{SL} = 0.013$ pF	$C_{SL} = 0.006$ pF	Ceq = 0.01 pF Cross sectional size for cross coupling capacitance = 50 x 50 mils, $\varepsilon_r = 2.38$, t = 60 mils
Lr = 1.013 nH, Cr =1 pF(Resonators)	Lr = 1.2059 nH, Cr =1 pF(Resonators)	Lr = 0.107825 nH, Cr = 10 pF(Resonators)	Cr = 0.35 pF, Q =1200 @5 GHz Short Circuit(SC_stub) = 280 mils, (Resonators)
	$L_{sh} = 76.394$ nH due to the Pi-conversion of negative –K inverter		

transmission zeros. The main coupling between parallel-coupled lines of combline filter is inductive. This inductive coupling and inductors, as impedance transformers, can be used to realize the extra TZ, with capacitive source-load cross coupling. An equivalent circuit of the cross-coupled combline filter is shown in Fig. 8.

However, the required capacitive coupling in the circuit of Fig. 3 (see Table I) is so weak that it is not easy to realize as a conventional capacitor. We propose the use of a dielectric block metalized on the top surface to realize an inhomogeneous transmission line to achieve the capacitive source-load coupling in the combline filter (Fig. 7). The advantage of this approach is that the required cross coupling value is realized by changing the substrate thickness, dielectric constant, top metal size, or length of the dielectric block. Connection wire side effects (lead inductance) can be avoided.

An approximate calculation for the overlapping conductor size of the cross coupled-transmission line block is determined by parallel-plate capacitance formula (Fig. 8). Values of the cross coupling block and other combline filter parameters are

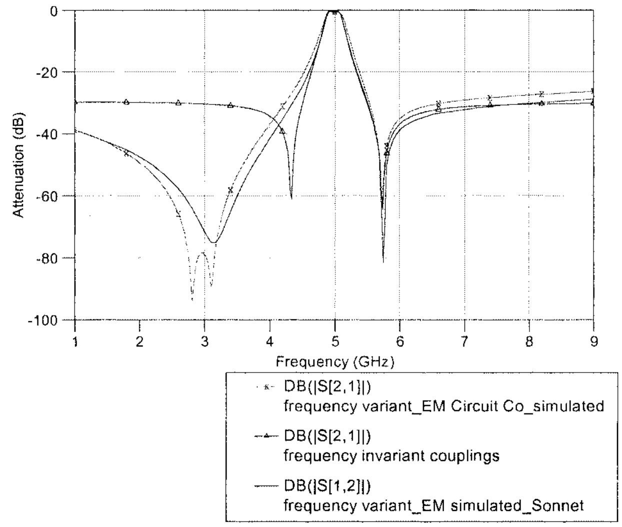

Fig. 9. Comparison of simulated filter response of the combline filter using a negative source-load coupling block.

displayed in Table I. Fig. 9 shows a comparison of the filter response for the frequency-invariant and frequency-variant couplings using Sonnet em® software. Even though the EM simulation result does not show two transmission zeros as clearly as the EM-circuit co-simulation result, it shows explicitly improvement in the lower stop bands when contrasted to the frequency-invariant couplings.

III. CONCLUSION

We have shown the difference between the filter response computed assuming frequency-invariant couplings from the real response based on real frequency-variant couplings. We propose to take advantage of the "real" couplings to predict and achieve extra transmission zeros by intentionally emphasizing the effect of the frequency variation. An efficient top metalized dielectric block overlay to be used in the planar bandpass filter design is presented for the source-load cross coupling.

REFERENCES

[1] R. J. Cameron, "General coupling matrix synthesis methods for chebyshev filtering functions," *IEEE Trans. Microwave Theory Tech.*, vol. 47, Apr. 1999.

[2] H. C. Bell, Jr., "Canonical asymmetric coupled-resonator filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, Sept. 1982.

[3] A. E. Atia and A. E. Williams, "Narrow-bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, Sept. 1972.

[4] R. Levy, "Direct synthesis of Cascaded Quadruplet (CQ) filters," *IEEE Trans. Microwave Theory Tech.*, vol. 43, Dec. 1995.

[5] S. Amari, "On the maximum number of finite transmission zeros of coupled resonator filters with a given topology," *IEEE Microwave Guided Wave Lett.*, vol. 9, Sept. 1999.

[6] ——, "Direct synthesis of folded symmetric resonator filters with source-load coupling," *IEEE Microwave Wireless Compon. Lett.*, vol. 11, June 2001.

[7] S. Amari and J. Boremann, "Using frequency-dependent coupling to generate finite attenuation poles in direct-coupled resonator bandpass filters," *IEEE Microwave Guided Wave Lett.*, vol. 9, Oct. 1999.

[8] R. V. Snyder, "Synthesis of filters with Frequency-Variant Couplings (FVC)," in *Proc. IEEE Int. Microwave Symp. Workshop*, May 2001.

[9] J. Dobrowolski, *Introduction to Computer Methods for Microwave Circuit Analysis and Design*. Boston, MA: Artech House, 1991.