Modelling of Wide Band Combline and Interdigital Filters

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Abstract—**Traditional TEM resonator theory based design methods cannot predict precisely bandwidth of interdigital and especially combline filters. Even twice wider bandwidths than assumed can be measured in combline filters. Known explanations of the effect are discussed and rejected. Coupled resonators are investigated and proper models of coupling are proposed. The difference between coupling coefficient definition in an ideal inverter and real inverter coupling is explained. A new explanation of bandwidth increase and method for design of wide band combline and interdigital filters are presented. The design examples of a two-pole 61% wide filter, three-pole 45 % wide filter and four-pole 49% wide filter are presented. The measurements show a very good agreement with theory.**

Keywords—**Microwave filters, bandpass filters, combline filters, interdigital filters, coupled resonators.**

I. Introduction

Traditional combline and interdigital filter theory [1]-[4] based on TEM mode coupling results in bandwidths that are too large taking into account the design assumptions. The bandwidth ratio defined as actual bandwidth to TEM bandwidth increases from 1 to over 2. The higher filter bandwidth the problem is getting worse especially in combline filters. Interdigital filters are less prone to bandwidth increase. The bandwidth increase was noticed in sixties [1] and seventies [2] of the previous century but attributed to inaccurate design data and "various approximations involved in the design equations" [1]. At the end of the previous century the problem was extensively examined and explained by several theories. Two of them are the most important. The effect of variable end loading of resonators [5], [6] and the evanescent waveguide mode additional coupling occurring when the ratio of filter cross-section height to wavelength is over 0.08 [7]. Certainly other explanations like influence of coupling between nonadjacent resonators is not a case [7] although the nonadjacent couplings have influence on the filter characteristics as will be shown later. But there are two more explanations of the problem. The bandwidth increase is due to misunderstanding of the coupled transmission line resonators behaviour and due to application of Cohn's theory for direct coupled filters [8] at least in a case of two pole filters. In traditional theory [1]-[4] the classical model of coupled lines is used which neglects the important information on the initial i.e. uncoupled lines state [8, 9] but it is not the reason for the design problems. The main problem is with interpretation of the electric length of resonators. The eigenfrequencies of coupled resonators are analyzed allowing proper selection of a model of coupled resonators. Proposed models of coupled combline and interdigital resonators involve mixed couplings i.e. electric and magnetic coupling simultaneously. The difference between coupling coefficient definitions in coupling structures including ideal inverters and real inverters is explained. A new interpretation of coupling phenomena is presented leading tothe accurate design of combline and interdigital filters. As an example of the method accuracy a 61 % wide two-pole, 48 % wide three-pole and 45 % wide four-pole combline filters have been designed, realized and measured. The measurement results are in good agreement with theory. The interdigital filters have been also designed and simulated in electromagnetic simulator.

It should be also noted that the design problems can be avoided by using 3-D electromagnetic simulators and optimization procedures but such a solution is really time consuming. The presented method can be much faster and sufficiently accurate. Moreover the method gives an interesting insight into coupling phenomena.

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II. Coupled Combline and Interdigital Resonators

Let us consider two coupled resonators of cross-section as shown in Fig.1. The combline and interdigital resonators have the same cross-section. The coupling implies existence of two resonant frequencies corresponding to the odd and even resonant mode. The higher the coupling between resonators the bigger is distance between resonant frequencies. From resonant frequencies the coupling coefficient can be computed as shown in [11]-[19].

$$k = \frac{\left|f_2^2 - f_1^2\right|}{f_1^2 + f_2^2} \tag{1}$$

The resonant frequencies of coupled combline and interdigital resonators have been computed versus the distance **d** between resonators and are presented in Fig.2. The 3D electromagnetic (QuickWave [20]) based on FDTD method has been used in these computations. The resonators are 30 mm long and have the loading capacitance setting the resonant frequency of a single uncoupled resonator at 1.27 GHz corresponding to electric length θ = 45.77˚. The resonators are square and have cross-section **s** 5x5 mm and are situated between metal walls separated by **b** = 15 mm, $\mathbf{s_0}$ = 5 mm. The end capacitance is realized with a metal tuning screw of 3 mm in diameter moving inside a circular hole of 4 mm in diameter. As one can see the resonant frequencies of modes are nearly the same when resonators are far from each other. When the distance between them approaches to zero resonant frequencies change in an asymmetric manner. The coupling coefficient between two resonators is also shown in Fig.2. In a case of combline resonators the resonant frequency of the odd mode changes much less than the resonant frequency of the even mode. Thus when a combline filter is built center frequency of the filter is shifted up from its initial value and the wider filter bandwidth the bigger is frequency shift. The frequency of the maximum of return loss (RL) characteristic in passband of the filter can be even approximately computed using the formula:

$$f_c = \frac{f_1 + f_2}{2} \tag{2}$$

E.g. the center frequency for the distance d = 0.3 mm is 1638 MHz and for distance d = 0.8 mm is 1501 MHz. The formula (2) is approximate only and especially in multiple resonator filters where the distances between resonators differ significantly the center frequency of the filter must be somehow averaged. In a case of interdigital filters both resonant frequencies change more uniformly form the frequency of uncoupled resonators and the center frequency of filters is easier to predict.

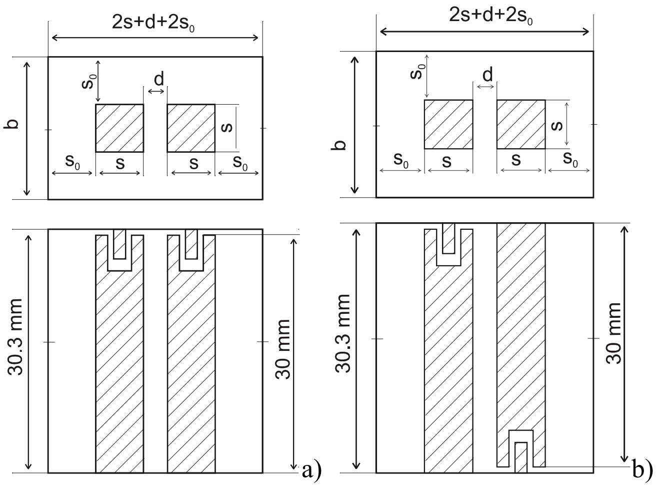


Fig.1. Cross-section of the combline (a) and interdigital (b) filter.

It is important to notice the difference between center frequency of the filter and resonant frequency of uncoupled resonators. In traditional design [1], [3] the electric length of resonators at filter center frequency is used. The formulas relating coupling coefficient between parallel coupled transmission lines with coupling between resonators of a certain electrical length are used:

combline filters:

$$K_{i,j}(\theta) = k_{i,j}\,\frac{\tan\theta\left(\theta\csc^2\theta + \cot\theta\right)}{2} \tag{3}$$

interdigital filters:

$$K_{i,j}(\theta) = k_{i,j}\,\frac{\sin\theta\left(\theta\csc^2\theta + \cot\theta\right)}{2} \tag{4}$$

where: θ is the resonator electrical length, $k_{i,j}$ is coupling coefficient between resonators obtained from synthesis i.e. from Cohn's theory [8].

The formula for combline filters is much more sensitive on electrical length than the formula for interdigital filters. Due to difference between center frequency of filters and resonant frequency of uncoupled resonators an application of formulas (3) and (4) leads to errors in coupling coefficient. And the error can be significant especially for combline filters with electrically long resonators. Thus when the electric length of resonators computed at filter center frequency is used the overcoupled resonators and wider filter bandwidths are obtained.

In general the problem of center frequency is quite interesting and can be attributed to the coupling phenomenon or more precisely to our possibility to model coupling between resonators. In general different inverters produce filters with different center frequencies in relation to passband edge frequencies. To consider the problem precisely the resonant frequencies of the coupled combline and interdigital resonators are compared with the resonant frequencies of the lumped shunt-type LC resonators coupled through J-inwerters [1] i.e. the series inductive mutual coupling and series capacitive mutual coupling. Formulas describing couplings are given below [11], [12], [19]:

-series capacitance mutual coupling

$$f_1^2 = f_0^2\,\frac{1}{1-k} \tag{5}$$

$$f_2^2 = f_0^2\,\frac{1}{1+k} \tag{6}$$

-series inductance mutual coupling

$$f_1^2 = f_0^2(1-k) \tag{7}$$

$$f_2^2 = f_0^2(1+k) \tag{8}$$

where: f_0 is the resonant frequency of uncoupled resonant circuits, and k is coupling coefficient which can be computed from resonant frequencies according to formula (1).

The resonant frequencies of resonators are normalized by the resonant frequency of uncoupled resonators and drawn versus the coupling coefficient in Fig.3 and Fig.4.

It is worth mentioning that the eigenfrequency equations relating resonant frequencies of coupled LC circuits to coupling coefficient were given long ago by Howe [11] and later repeated by Sturley [12]. Over twenty years ago the coupling coefficient was computed from the resonant frequencies of dielectric resonators [13] what started the eigenfrequency method so popular now [14]-[19], [21]-[22].

As can be seen in Fig. 3 the coupling between combline resonators is such that there is not easy to apply a proper model. In fact a mixed coupling model [12], [14], [17] should be used. In this model two J-inverters of the series capacitance mutual coupling type and of series inductance mutual coupling type are used simultaneously. The total coupling coefficient depends on both capacitive and inductive coupling. Only for filter bandwidths below 10% (coupling coefficient below 0.1 approximately) resonant frequencies of coupled combline resonators behave according to series capacitance coupling model. Wider filters have center frequencies shifted up significantly in comparison with capacitance coupling model.

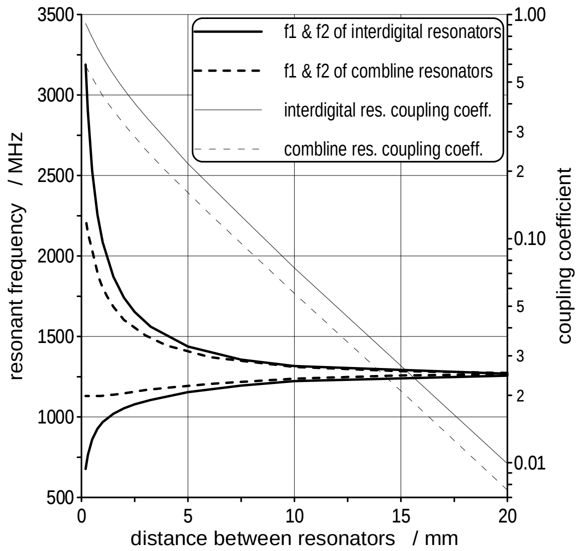


Fig.2. Resonant frequencies of coupled combline and interdigital resonators and coupling coefficient versus distance between resonators.

As can be seen in Fig. 4 the coupling between interdigital resonators can be quite well described with series capacitance coupling when they are ~46º long but for the coupling coefficient not exceeding 0.6.

The series inductance mutual coupling quite precisely approximates coupled interdigital resonators for any coupling coefficient when they are ~82º long. The different types of coupling imply different formulas for center frequencies of the filters built with interdigital resonators, which are electrically short (well below 60º) or long (well over 60º). Proper formulas for center frequencies can be found in [1].

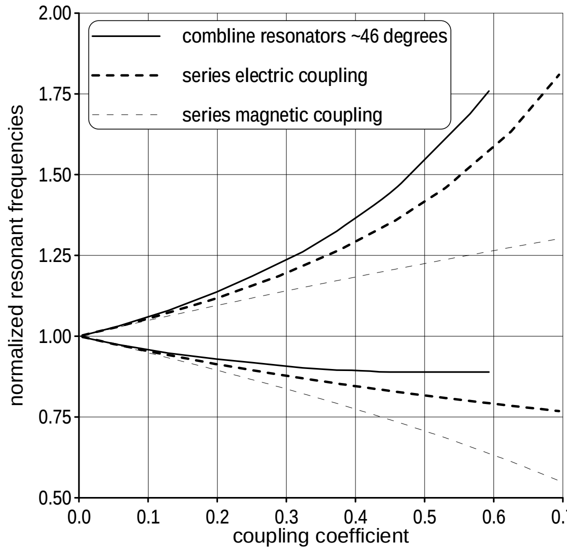


Fig.3. Resonant frequencies of combline resonators compared with resonant frequencies of models of series coupling.

Although for the resonator approximately 60º long the center frequency of interdigital filters is difficult to predict. Presumably design of interdigital filters employing resonators having length close to 90º can be quite precise when based on series inductance coupling model. But the model cannot predict precisely the out of band characteristics. For shorter resonators the situation can be unclear if the filter bandwiths are wider than 60%. The center frequency will be lower than predicted from capacitance coupling model. The mixed coupling model seems to be the most accurate also in a case of interdigital resonators.

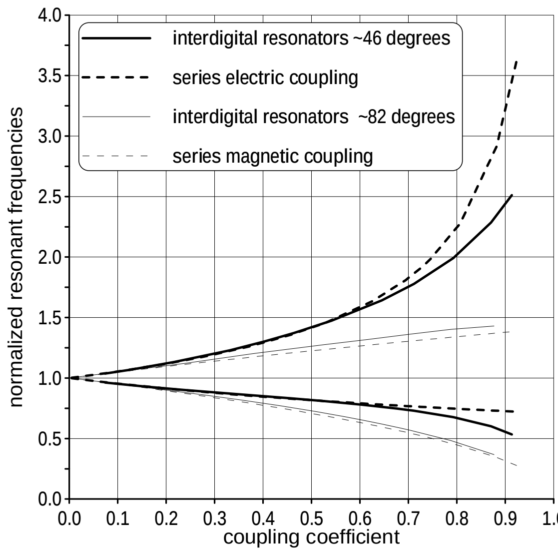


Fig.4. Resonant frequencies of interdigital resonators compared with resonant frequencies of models of series coupling.

Formulas describing mixed couplings shown in Fig.5 are given below [18]-19]:

-series capacitance and series inductance mutual coupling (the same sign of coupling elements L_s and C_s), k_L=L/L_s, k_C=C_s/C

$$f_1^2 = f_0^2 \frac{1-k_L}{1-k_C} \tag{9}$$

$$f_2^2 = f_0^2 \frac{1+k_L}{1+k_C} \tag{10}$$

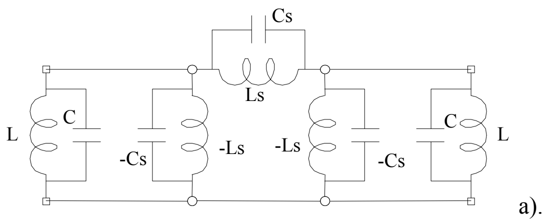
$$k = \frac{f_2^2 - f_1^2}{f_1^2 + f_2^2} = \frac{k_L - k_C}{1 - k_L k_C} \tag{11}$$

-series capacitance and series inductance mutual coupling (different signs of coupling elements), k_L=L/L_s, k_C=C_s/C

$$f_1^2 = f_0^2 \frac{1-k_L}{1+k_C} \tag{12}$$

$$f_2^2 = f_0^2 \frac{1+k_L}{1-k_C} \tag{13}$$

$$k = \frac{f_2^2 - f_1^2}{f_1^2 + f_2^2} = \frac{k_L + k_C}{1 + k_L k_C} \tag{14}$$



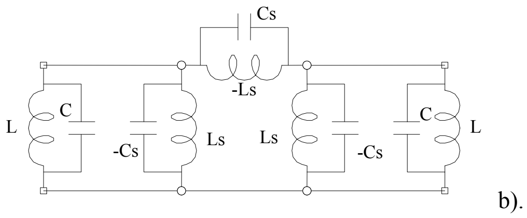


Fig.5. Series capacitance and inductance coupling models:
a) with coupling elements of the same sign,
b) with coupling elements of different sign.

Results show that proper real inverters should be used to model combline and interdigital resonator couplings. In a case of combline resonators mixed series coupling model should be used with both positive coupling elements. Such model preserves the transmission zero at frequency corresponding to electric length of 90 degrees. In a case of interdigital resonators the situation is much more complicated. The series magnetic coupling model could be suitable for electrically long resonator and the series electric coupling model could be suitable for electrically short resonators but they are good in prediction of center frequency only. The transmission characteristic can be predicted more precisely with the mixed coupling model. The mixed series coupling model with coupling elements of different signs is the most suitable for interdigital resonators of any electric length.

III. Ideal Inverters versus Real Inverters

The traditional design method is based on the ideal inverters what is another source of problems. Coupling coefficient between resonators in combline and interdigital filters are computed on the base of Cohn's method [1,8]. This can be a source of design errors especially when the results of the eigenfrequency method are used to find couplings and distances between resonators. Cohn used ideal inverters to develop formulas for coupling coefficients. In the eigenfrequency method the real inverters are used. The coupling coefficient in Cohn's method has different definition that the coupling coefficient in the eigenfrequency method. When two resonant circuits coupled with ideal inverters are considered the coupling coefficient (k_{Cohn}) depends on its two eigenfrequencies as follows:

$$k_{Cohn} = \frac{f_2 - f_1}{f_0} \tag{15}$$

where: f_0 is the resonant frequency of uncoupled resonators and simultaneously the center frequency of the filter

$$f_0 = \sqrt{f_1 f_2} \tag{16}$$

The formula (15) is obviously different than (1) and the coupling coefficient k_{Cohn} is different than k. Difference is quite small for weak couplings and can be quite big for strong couplings. In majority cases the following approximate formula can be used to transform the "ideal" Cohn's coupling coefficient to the "real" coupling coefficient used in the eigenfrequency method:

$$k = \frac{k_{Cohn}\sqrt{k_{Cohn}^2 + 4}}{k_{Cohn}^2 + 2} \tag{17}$$

The "real" coupling coefficient is always smaller than "ideal" coupling coefficient and always smaller than 1. The "ideal" coupling coefficient k_{Cohn} can be much bigger than 1. When the coupling coefficients in a filter are calculated according to the direct-coupled resonator filters theory and the geometry of the filter is established on the base of "real" coupling coefficients computed with the eigenfrequency method the resulting filter can be wider than expected.

IV. Ground Plane Spacing Influence on Combline Filters

To check the influence of the ground plane spacing **b** and of variable end loading of resonators on coupling between combline resonators the coupling coefficient versus distance **d** has been computed for many different structures. The eigenfrequency method as described above has been used. It is enough to compare the following three structures having the cross-section as in Fig.1 with the following parameters:

1. **b** = 15 mm, **s** = 5 mm, $\mathbf{s_0}$ = 5 mm, **l** = 30 mm (resonator length), f(uncoupled) = 1.27 GHz, **b/λ** = 0.0852
2. **b** = 30 mm, **s** = 10 mm, $\mathbf{s_0}$ = 10 mm, **l** = 30 mm (resonator length), f(uncoupled) = 1.27 GHz, **b/λ** = 0.1704
3. **b** = 30 mm, **s** = 5 mm, $\mathbf{s_0}$ = 5 mm, **l** = 30 mm (resonator length), f(uncoupled) = 1.123 GHz, **b/λ** = 0.1704

The second case is exactly twice bigger in cross-section as the first one and has different loading capacitance only to keep the same resonant frequency of uncoupled resonators. The third case and first one have the same loading capacitances what gives different resonant frequencies. The results of computations are shown in Fig. 6. As one can see the values of coupling coefficient do not differ significantly (in fact they are nearly the same) in the area of strong coupling thus the ground plane spacing influence explanation [7] cannot longer be valid. As well as the variable end loading [5]-[6] explanation is not true to some extends. The evanescent modes should have much more visible influence for small distances between resonators than for big ones. Thus when the distance between resonators is big and the coupling coefficient in structure **2** is 10% bigger than in structure **1** it is not the evanescent mode effect. The difference between structure **1** and **2** seen in Fig.5 results from different end loading. But clearly the effect cannot be a reason for 100% bandwidth expansion. The difference between structures **1** and **3** results from different impedance of lines. The area of weak couplings is also interesting because it shows that dependence of the coupling coefficient on electrical length of lines as stated in TEM theory [1], [3] has limited accuracy and formulas (3) and (4) cannot be considered as exact.

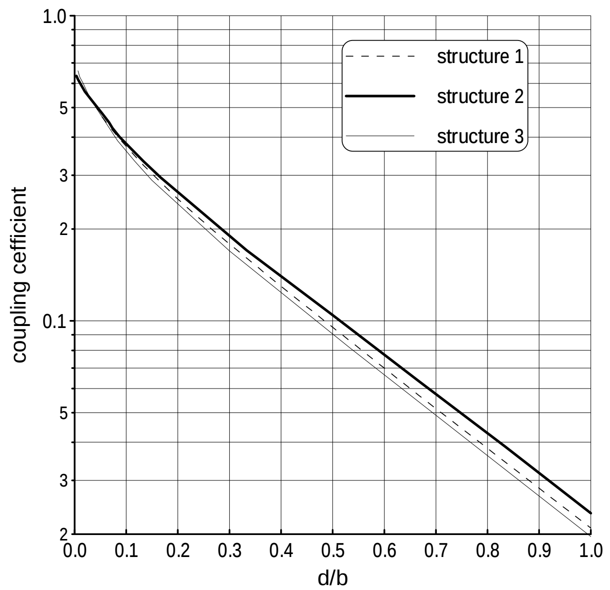


Fig.6. Coupling coefficient between two coupled combline resonators in different structures.

V. INFLUENCE OF NONADJACENT COUPLINGS

In wide band combline and interdigital filters the nonadjacent couplings have quite significant values. Their presence is easily observed due to multiple transmission zeros at frequencies above the filter passband or reduction of transmission zeros. The nonadjacent couplings can be also analyzed by means of the eigenfrequency method as in [14,19], where a case of three coupled resonators is described. The nonadjacent couplings are spurious but should be taken into account in the design process. For example in a case of three-pole combline filter shown below the spurious coupling between resonators 1 and 3 changes the couplings between resonators 1 - 2 and 2 - 3. The change is such that the couplings 1 - 2 and 2 - 3 should be increased to obtain assumed filter bandwidth. The nonadjacent coupling between resonators 1 and 3 results in 15% increase of the couplings between resonators 1 - 2 and 2 - 3. Because the couplings should be increased the nonadjacent couplings cannot be a reason for too wide filter bandwidths. Unfortunately nonadjacent couplings cannot be avoided. The design method should take them into account. Thus the traditional method based on direct- coupled resonator filter theory is not sufficient. In fact the electromagnetic simulations are needed and cannot be avoided.

VI. DESIGN OF WIDE BAND FILTERS

The filter bandwidth according to tradition is as follows:

$$w = \frac{f_{p2} - f_{p1}}{f_{ca}} \tag{18}$$

where: f_{p1} and f_{p2} are passband edge frequencies, f_{ca} is the center frequency of the filter taken arithmetically in the middle between f_{p1} and f_{p2} .

The definition of center frequency is arbitrary and is not related to the shape of the return loss (RL) characteristic. Different types of filters with the same passband edge frequencies have different shapes of the RL characteristic and different distribution of frequencies of minima and maxima in the passband. The formula (2) predicts the frequency of maximum of the RL characteristic in the passband.

In the first step the coupling coefficients between resonators $k_{i,j}$ should be calculated for given filter. The traditional method [1,3] can be used to calculate coupling coefficicients approximately. According to Cohn [8]:

$$k_{Cohn(i,j)} = \frac{w}{\sqrt{a_i a_j}} \tag{19}$$

where: $\boldsymbol{w}$ is filter bandwidth, $\boldsymbol{a_i}$, $\boldsymbol{a_j}$ normalized elements of the low pass prototype filter.

The "ideal" coupling coefficients obtained from equation (19) should be transformed to "real" coupling coefficients using formula (17).

It is not enough to assume that each resonator is coupled with two other nearest resonators. The "real" coupling coefficients are just starting point to the synthesis or optimization method for multiple coupled resonator filters. In a case of synthesis the values of couplings between nonadjacent resonators will probably be not correctly related to the physical structure. For optimization one has to know couplings between nonadjacent resonators thus optimization should use data from electromagnetic analysis e.g. LINPAR [23] or any 3D electromagnetic simulator. The external quality factors Q_{ext1} and Q_{extn} should also be obtained from synthesis procedure.

In a case of two-pole filters the "real" coupling coefficient is what one needs to realize a filter except Q_{ext} which can be easily find by optimization in a circuit simulator. To optimize the combline filter the circuit shown in Fig.7a can be used while the interdigital filter can be modelled with circuit shown in Fig.7b. In Fig.7a the resonator consists of the parallel resonant circuit LC corresponding to a transmission line of certain electrical length and the loading capacitance C^e. The coupling structure is realized with real inverter of mixed type as in Fig.5. The lines of impedance Z and length l represent input tap lines. The LC circuit has the resonant frequency set at the frequency corresponding to 90 degrees electric length of the combline resonator. As one can see elements L_s and C_s create a parallel resonant circuit that has the same resonant frequency. Thus at the frequency corresponding to 90 degrees there is no coupling. The coupling between LC circuits corresponds to the coupled transmission lines thus the capacitive coupling coefficient and inductive coupling coefficient between LC resonant circuits are the same (k_L = L/L_s = k_C = C_s/C). The presence of loading capacitance C^e decreases the value of total capacitive coupling k_{Ctotal} = $C_s/(C+C^e)$. Consequently the total coupling coefficient of the structure in Fig.7a is as follows:

$$k_{total} = \frac{k_L - k_{Ctotal}}{1 - k_L k_{Ctotal}} \quad (20)$$

In a similar manner higher order combline filters can be modelled including couplings between nonadjacent resonators. The main and cross couplings levels can be adjusted using a tool like LINPAR [23].

The circuit of the interdigital filter is created in the same manner except the different signs of the coupling elements. The total coupling coefficient of the structure in Fig.7b is given by the following equation:

$$k_{total} = \frac{k_L + k_{Ctotal}}{1 + k_L k_{Ctotal}} \quad (21)$$

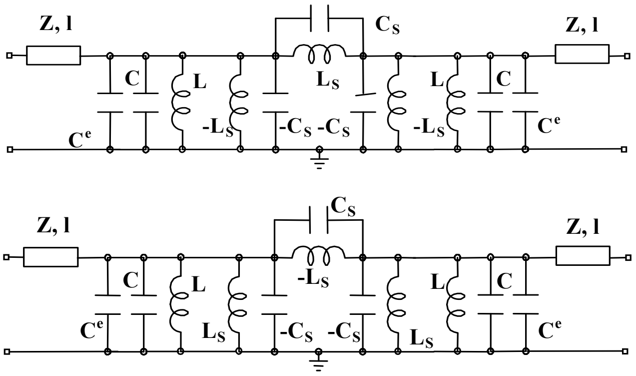


Fig. 7. a). Equivalent circuit of a combline two-pole filter.
b). Equivalent circuit of an interdigital two-pole filter.

When such circuits for higher order interdigital filters are created one should remember that resonators 1 and 3 or 2 and 4 are oriented as in combline filters thus the real inverters with the elements of the same signs should be used to model coupling between them.

The coupling coefficients between resonators can be transformed into geometry i.e. distances between the lines using data as in Fig.2 or even lookup tables or graphs [24] but in the latter the accuracy can be decreased due to limited accuracy of formulas (3) and (4). The transfer of external quality factors into geometry is more demanding and use of electromagnetic simulator is recommended.

After that the only unknowns are loading capacitances C^e. Their approximate values can be computed using formula [1]:

$$C^e = \frac{1}{\omega Z} \cot \theta \quad (22)$$

where: $\boldsymbol{\theta}$ is the electrical length at the resonant frequency of the uncoupled resonator, $\boldsymbol{Z}$ is resonator impedance, $\boldsymbol{\omega}$ is the angular frequency of uncoupled resonators.

Loading capacitances depend on the resonator position. In combline filters inner resonators have bigger loading capacitances than computed according to formula (22) in interdigital filters inner resonators have loading capacitances smaller than computed from formula (22).

VII. Simulation And Experimental Results

The example filters of two, three and four resonators have been designed and realized. Two-pole filters have been designed as described above. Three and four resonator filters have been designed by optimization in QuickWave. Assumed bandwidths were 61 % for 2-pole filter, 45% for 3-pole filter and 50 % for 4-pole one. The types of characteristics: n = 2 equiripple with ripples 0.36 dB (RL = 11 dB), n = 3 equiripple with ripples 0.0346 dB (RL = 21 dB) and n=4 equiripple with ripples 0.0346 dB (RL = 21 dB). All filters should have the same center frequency of 1.5 GHz. In Table 1 the “ideal” coupling coefficients calculated from (19) and transformed to “real” coupling coefficients (using (17)) are compared with coupling coefficients realized in filters. The external quality factors and distances between resonators are also shown in Table 1. Distances between resonators have been approximated slightly in order to apply a precise gap gauge.

Filters have been realized using resonators of square cross-section 5x5 mm. The electrical length of resonators has been chosen as 45.77º at 1.27 GHz (physical length 30 mm). The housings have inner dimensions 15 x 20.3 x 30.3 mm, 15 x 26 x 30.3 mm and 15 x 32 x 30.3 mm respectively. N-type connectors have been used at input and output and their inner connectors have been soldered to the resonators at position 25.6 mm from the resonator ground point preserving the same external quality factors for all filters. The resonators have been tuned by means of metal screws with diameter of 3 mm moving inside resonators (holes of 4 mm in diameter). The tuning screws have been used to realize the loading capacitances C^e as well as preserve the symmetry of characteristics. It is worth noting that the C^e in two-pole filter has value corresponding to 1.27 GHz and realizes the filter with center frequency $\boldsymbol{f_{ca}}$ = 1.5 GHz. The frequency of the maximum of RL in the passband is 1.61 GHz what is in a quite good agreement with (2). Filters have been made of brass. Housings have been silverplated while resonators have not. A photograph of two-pole filter is shown in Fig. 8. Results of measurements are shown in Fig. 9, Fig. 10 and Fig. 11.

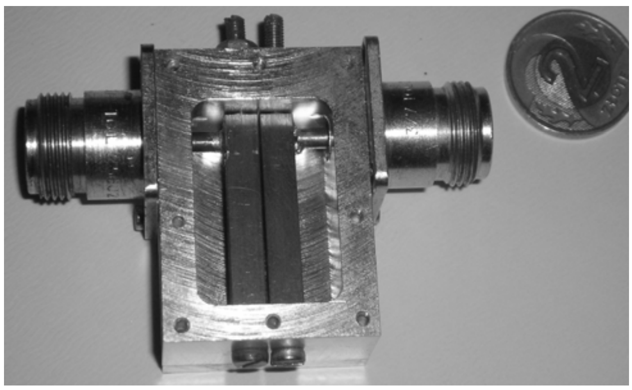


Fig. 9. Photograph of the two-pole combline filter without cover.

The two-pole filter has 916 MHz wide bandwidth and RL level 10.7 dB, which corresponds quite well to assumptions. The center frequency is 1500 MHz. Thus bandwidth is exactly

61 %. The situation would be different if the traditional design method were used. The center frequency of the filter and Cohn's coupling coefficient used to calculate the electrical length of the resonators and then coupling between them would produce 30 % bigger coupling coefficient $K_{1,2}$ what means that the filter bandwidth would be significantly increased and match deteriorated.

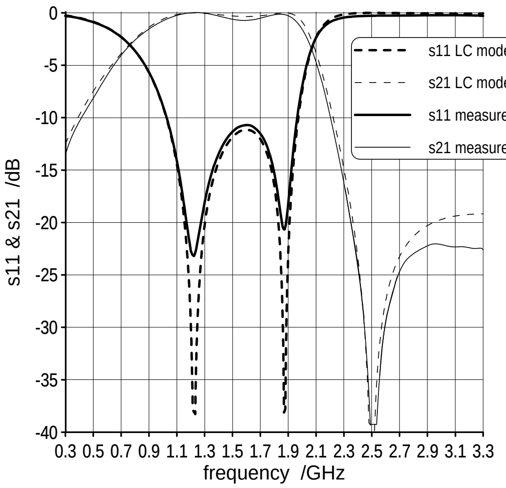


Fig. 9. Measured characteristics of the 61% wide two-pole combline filter compared with the characteristics of the lumped element model (L =1.52 nH, C =2.666 pF, L_s=1.951 nH, C_s=2.077 pF, C^e=2.73 pF, Z =85 Ω, l = 5 mm).

The three-pole filter has bandwidth of 673 MHz and RL level 20.5 dB, which again corresponds well to assumptions. The center frequency is 1499 MHz. Thus bandwidth is 44.9 %. The four-pole filter has bandwidth of 726 MHz and RL level better than 21 dB, which again corresponds well to assumptions. The center frequency is 1495 MHz. Thus bandwidth is 48.6 %.

TABLE 1.
COUPLING COEFFCIENTS, EXTERNAL QUALITY FACTORS AND DISTANCES BETWEEN RESONATORS.

Filter Type	filter order	k_{Cohn}	k Eq.17	k realized	Q_{ext_Cohn}	Q_{ext} realized	d /mm
Comb.	n=2	k_{12}= 0.6505	0.5646	0.56	2.054	2.1	0.3
	n=3	k_{12}=k_{23}= 0.4795	0.4422	0.52	1.804	2.1	0.5
	n=4	k_{12}=k_{34}= 0.4674	0.4327	0.50	1.784	2.1	0.6
		k_{23}= 0.3565	0.3405	0.46			0.8
Int.	n=2	k_{12}= 0.6965	0.5936	0.6	1.5873	2.1	1.15

A two-pole interdigital filter has been designed and simulated in QuickWave [20]. The filter has 42 % bandwidth at center frequency 1.59 GHz and ripple level in the passband 0.0436 dB (RL = 20 dB). The computed characteristics are shown in Fig. 12. The resonators of the same cross-section and length as in combline filters have been used. The input/output structure has also been the same.

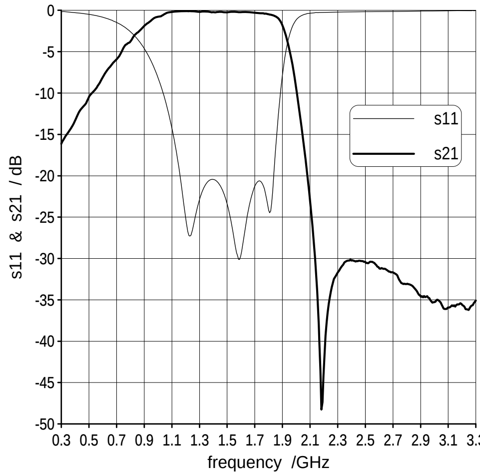


Fig. 10. Measured characteristics of the 45% wide three-pole combline filter.

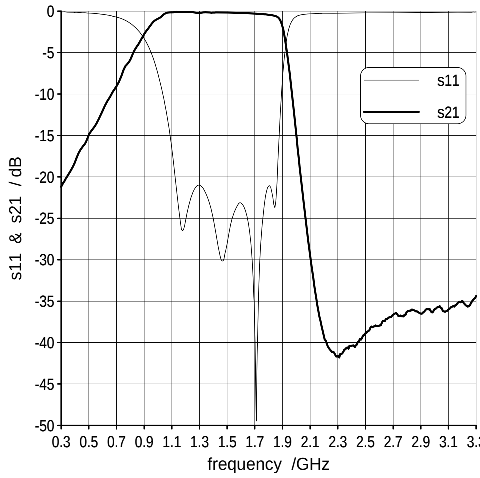


Fig. 11. Measured characteristics of the 49% wide four-pole combline filter.

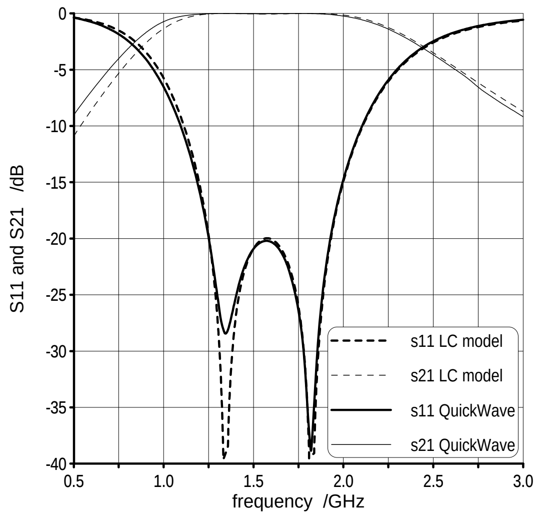


Fig. 12. Simulated characteristics of the 61% wide two-pole interdigital filter compared with the characteristics of the lumped element model (L =3.092 nH, C =1.3108 pF, L_s=6.6681 nH, C_s=0.6078 pF, C^e=2.117 pF, Z =85 Ω, l = 5 mm).

The coupling coefficients compared in Table 1 confirm ideas given in previous paragraphs of the paper and justify the proposed design method. Except two-pole filters, the realized coupling coefficients are bigger than calculated from direct-coupled resonator filters theory [8]. Thus the bandwidth increase is attributed to the wrong definition of electrical length of resonators. The resonant frequency of uncoupled resonators is the frequency that scales the coupling between them thus the coupling parameters should be computed at the frequency of the uncoupled resonators. The bigger coupling coefficients are due to nonadjacent resonator couplings. Two-pole filters need correction of the coupling coefficient calculated according to [8] as given by Eq.17. In any case the external quality factors must be adjusted. The design of higher order filters can be hardly done without electromagnetic simulations.

VIII. CONCLUSION

A new explanation of the bandwidth expansion effect in combline and interdigital filters has been presented. Coupled resonators have been analyzed by means of the eigenfrequency method. The results enable to reject known explanation of the bandwidth expansion effect and justify the change of definition of the resonator electrical length. Moreover proper coupling models of combline and interdigital resonators are presented. The difference between coupling coefficients used in the traditional design and in the eigenfrequency method is shown and method to overcome the difference is given. The trial combline filters of 62%, 45% and 49 % wide bandwidths have been realized and measured. The interdigital filter has been also simulated. The results of the experiments justify presented approach thus the discrepancy between traditionally designed and realized filter bandwidths has been explained as resulting from the misunderstanding of the coupling mechanism. The design of combline and interdigital filters based on the traditional method should be limited to double resonator filters or in a case of higher filter orders to narrow bandwidths.

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