Constraints to the Optimum Performance and Bandwidth Limitations of Diplexers Employing Symmetric Three-Port Junctions

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***Abstract*—This work addresses the problem of the fundamental limitations to the optimum performance of diplexers employing three-port junctions and preassigned branching filters. In this situation it is a common misconception that optimum results are achieved by utilizing a good power divider closed by two good branching filters. From the properties of the S-matrix of a three-port junction, we show this not to be the case and derive a set of necessary conditions to be satisfied in order that the junction be successfully employed in the realization of diplexers. We derive explicit expressions for the positions at which the filters must be placed in the junction arms for optimum diplexer performance, resulting in considerable simplification of the difficult diplexer synthesis problem. A theorem on the maximum achievable bandwidth is proved and validated by means of two practical examples.**

I. Introduction

AT MILLIMETER frequencies it is expedient to employ reciprocal circuits for the realization of diplexers and multiplexers and some interesting examples have appeared in the recent literature [1]–[4]. These are constituted essentially by three-port junctions closed by two filters, which select the RX signal at port 1 and the TX signal at port 2, while port 3 is the input port in Fig. 1. The junction must be realized in such a way that port 3 and 1, 3, and 2 be perfectly matched over the RX band and the TX band, respectively.

A general theory for the synthesis of diplexers-multiplexers is indeed presented in [5]–[7]. However, its application is not immediate since the deviations between prototype and actual physical structure require a further process of optimization. Therefore, for the practical purpose of design, the geometry of the junction, including the distances from the junction at which the filters need to be positioned, is often determined by making use of either of the two following approaches [8].

1) Separate design of the three-port junction, of the branching filters and selection of the optimum distances of the filters from the junction.
2) Global optimization of the whole diplexer by varying either filter dimensions or junction geometry and filter distances or both together.

The latter approach may seem more satisfactory, but, in practice is rendered problematic by the large number of variables and by the "error function" not being analytically available, so that a number of local minima occur.

With a view to bypassing such difficulties, some researchers have employed segmentation techniques (evolution strategy method) in the solution space, that while yielding some very good results [1], [2], appear to be computationally cumbersome. The first approach would seem more expedient, but, unfortunately, its results so far are disappointing until and unless either the filters, the junction or both are altered essentially, falling back therefore into case 2) above.

This work examines the reasons for this failure starting from the properties of the S-matrix of the lossless reciprocal three-port junction. It is shown that, contrary to common belief, the ideal junction is not provided by a matched power splitter and that, in fact, it has to satisfy some necessary conditions. These will be derived and illustrated in the following by means of two examples of diplexer designs, the first one from the literature, the second a new one.

Once the junction is well defined, we obtain by means of a simple analytical formula the distances of the filters from the junction for optimum performance.

Moreover, there is a direct design application of the above criteria: often, in fact, one has to assemble a diplexer starting from two given filters and a three-port junction, with just the possibility of experimental tuning. In this case, the analytical results here obtained permit to simplify considerably the procedure, concentrating attention just upon the junction, whose characteristics can be modified in order to satisfy simple prescribed specifications. The two filters are then connected to the junction at distances determined from separate measurements of junction and filters.

It is noted that, for the sake of simplicity, the analytical model is developed on the basis of just fundamental mode interaction between junction and filters. In cases where optimum distances are so short that higher modes may cause interaction, the single mode result may be considered as an excellent starting point for an optimization routine anyway. It is emphasized that said criteria are based on the analysis of the scattering matrix of the junction and they are, as such, of general application, independently of the technology adopted.

II. Analysis

In Fig. 1, the diplexer is modeled as a lossless reciprocal three-port junction (J) with ports 1 and 2 closed by filters $F1$

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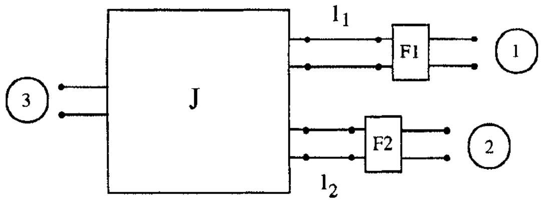


Fig. 1. Black box equivalent circuit of the diplexer; J is a three-port junction, $F1$ and $F2$ are the two filters.

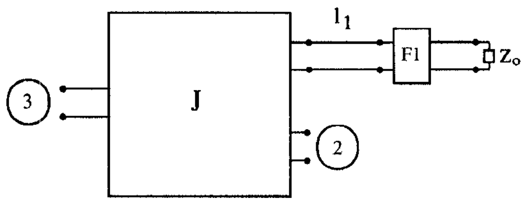


Fig. 2. Two-port network obtained by closing port 1 of J by $F1$, located at distance l_1 from the junction.

and $F2$ with passbands b_1 and b_2, respectively. We require:

1) perfect transmission between ports 3 and 2 over the band b_2;
2) perfect transmission between ports 3 and 1 over the band b_1.

Let us consider the first requirement:

This is certainly satisfied, if by closing port 1 by filter $F1$, terminated by a matched load (Fig. 2), the resulting two-port junction is perfectly matched and lossless over the band b_2.

Let ρ_{L1} denote the reflection coefficient of $F1$, located at a distance l_1 from port 1 of the junction, the scattering matrix of the resulting 2-port network (3-2) is given by

$$\begin{bmatrix} S_{22} & S_{23} \\ S_{23} & S_{33} \end{bmatrix} = \begin{bmatrix} s_{22} & s_{23} \\ s_{23} & s_{33} \end{bmatrix} + \frac{1}{\frac{1}{\rho_{L1} e^{-j2vl_1/a}} - s_{11}} \times \begin{bmatrix} s_{12}^2 & s_{12}s_{13} \\ s_{12}s_{13} & s_{13}^2 \end{bmatrix}. \quad (1)$$

Where s_{ij} are the parameters of the scattering matrix $\mathbf{S}$ of the junction J and $v = \beta a$ is the effective frequency, β being the propagation constant of the fundamental mode of the waveguide feeds and a being an appropriate transverse dimension. Requirement 1) therefore implies

$$S_{22}(v) = 0 \qquad \text{(over the band } b_2\text{)} \quad (2)$$

Hence, using the fact that for a lossless junction $s_{33}^* = \frac{s_{11}s_{22} - s_{12}^2}{\Delta s}$, where Δs is the determinant of $\mathbf{S}$, we derive

$$\rho_{L1} e^{-j2vl_1/a} = \frac{s_{22}}{s_{11}s_{22} - s_{12}^2} = \frac{s_{22}}{\Delta s\, s_{33}^*}. \quad (3)$$

Since $F1$ operates in its stopband over band b_2, we have $|\rho_{L1}| \cong 1$ to a good approximation, $|\Delta s| = 1$ due to losslessness and consequently from (3), in order for constraint i) to be satisfied, we must require

$$|s_{22}| = |s_{33}| \qquad \text{over the band } b_2. \quad (4a)$$

In analogous manner, we may show that in order for requirement 2) to be satisfied, we must have

$$|s_{11}| = |s_{33}| \qquad \text{over the band } b_1 \quad (4b)$$

It is noted that (4a) and (4b) above involve just the scattering parameters of the junction independently of the filters and of their spacings from the junction. In order that requirement 1) be satisfied at the frequency f_2, we choose the distance l_1 by solving (3) at $f = f_2$, yielding the spacing

$$l_1 = -\frac{a}{2jv} \ln \left[\frac{s_{22}}{\Delta s\, s_{33}^* \rho_{L1}} \right] \qquad f = f_2. \quad (5a)$$

Proceeding in analogous fashion for ports 2–3, we recover

$$l_2 = -\frac{a}{2jv} \ln \left[\frac{s_{11}}{\Delta s\, s_{33}^* \rho_{L2}} \right] \qquad f = f_1. \quad (5b)$$

The spacings l_1, l_2, obtained from (5a) and (5b), together with (4a) and (4b), ensure ideal diplexer behavior at the midband frequencies f_1 and f_2 of the two filters.

We shall now consider the question of the maximum achievable bandwidth under the constraint of junction symmetry.

A. Symmetrical Junctions

Let us now consider the important class of three-port junctions endowed with a plane of symmetry. To this class belong many common junctions, e.g., considering only rectangular waveguide junctions, the abrupt E-plane and H-plane junctions, T-junctions, Y-junctions with arbitrary angle of aperture and junctions such as those described in [2], [9], and [10] containing more complicated matching sections.

The above symmetry implies [11]

$$s_{11} = s_{22} \qquad \text{for } E\text{-plane junctions} \quad (6a)$$

$$s_{11} = -s_{22} \qquad \text{for } H\text{-plane junctions.} \quad (6b)$$

Consequently, by combining property (6) above with constraint (4a)–(4b), we deduce that the geometry is to be selected in such a way that

$$|s_{11}| = |s_{22}| \cong |s_{33}| \quad (7)$$

over the bands b_1 and b_2.

We may now state the following lemma:

Lemma: A necessary condition in order for (7) to be satisfied is that

$$|s_{33}| \geq 1/3. \quad (8)$$

Proof: We will first deal with E-plane junctions.

The above statement can be proved by means of geometrical considerations. In fact, define by Γ_e, Γ_o the reflection coefficients at port 1 corresponding to an even, odd excitation respectively at ports 1 and 2. We have then

$$s_{11} = \frac{\Gamma_e + \Gamma_o}{2} \quad (9a)$$

$$s_{12} = \frac{\Gamma_e - \Gamma_o}{2}. \quad (9b)$$

Because of losslessness we have $s_{33} = -\frac{s_{13}}{s_{13}^*}(s_{11}^* + s_{12}^*)$, so that

$$|s_{33}| = |\Gamma_e|. \quad (9c)$$

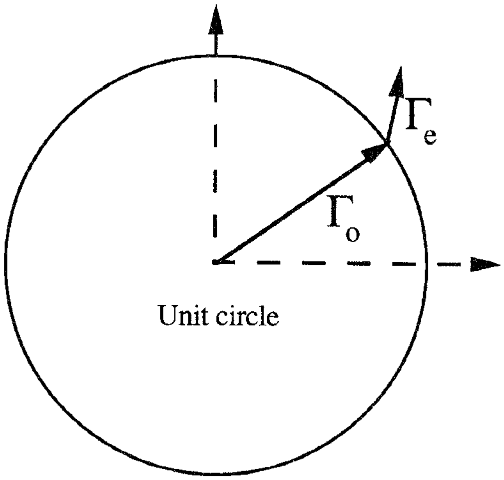


Fig. 3. Representation of the reflection coefficients Γ_e and Γ_o, corresponding to even/odd excitation, in the complex plane.

The condition $|s_{22}| = |s_{33}|$ over the band b_2 (4a) requires, however

$$|\Gamma_e| = \left|\frac{\Gamma_e + \Gamma_o}{2}\right|. \tag{10}$$

Let us plot the reflection coefficients as vectors in the complex plane (see Fig. 3). By virtue of the assumed symmetry, in the odd excitation case, the amplitude of the wave transmitted to port 3, $b_3 = s_{13}(\frac{1}{2}) + s_{13}(-\frac{1}{2}) = 0$, hence $|\Gamma_o| = 1$.

Noting, moreover, that the minimum of s_{11} is obtained if Γ_e and Γ_o have opposite phases, we obtain

$$\left|\frac{\Gamma_e + \Gamma_o}{2}\right| \geq \frac{1 - |\Gamma_e|}{2} \tag{11}$$

which together with (10) proves the statement (8).

The H-plane case follows simply by exchanging Γ_e with Γ_o. □

A direct consequence is that an optimum junction behavior for the purpose of realizing a diplexer is rather different from that of a power splitter, for which $|s_{33}| = 0$. We shall now consider the problem of diplexer bandwidth. Although the actual achievable bandwidth depends on the specific junction and filter characteristics, we will determine an upper limit to its value with the help of the following theorem.

Theorem: The maximum achievable bandwidth, for a given a reflection ε, of the two-port junction obtained by closing one arm of a symmetrical three-port on a given filter is given by

$$2\Delta v \leq \frac{1 - |s_{33}|^2}{|s_{33}|^2}\frac{\varepsilon}{2\bar{l}}$$

where s_{33} is the reflection at the common port of the junction, $\bar{l}$ is an "effective" normalized distance at which the filter is positioned.

Proof: By closing port 1 of the junction on $F1$ (Fig. 2), the reflection coefficient of the resulting two-port network $S_{22}(v)$ is expressed in terms of the s-parameters of the junction J and of $F1$ as

$$S_{22}(v) = s_{22}(v) + \frac{s_{12}^2(v)}{\frac{1}{\rho_{L1}e^{-j2vl_1/a}} - s_{11}(v)}. \tag{12}$$

We have seen that, according to (11), $S_{22}(v_2) = 0$. Therefore in a neighborhood of the midband frequency f_2, corresponding to v_2, considering the first order Taylor approximation to S_{22} we have

$$S_{22}(v) \approx \frac{d}{dv}S_{22}(v_2)\Delta v. \tag{13}$$

Now, we require that $|S_{22}(v)| < \varepsilon$ over the band Δv, ε being the maximum reflection coefficient of the junction that is compatible with the diplexer specifications, that is

$$\left|\frac{d}{dv}S_{22}(v_2)\right| \leq \frac{\varepsilon}{\Delta v}. \tag{14}$$

By differentiating (12) with respect v, we obtain

$$\begin{aligned}\frac{d}{dv}&S_{22}(v_2)\\ &= s'_{22} + \frac{2s'_{12}s_{12}}{\frac{1}{\rho_{L1}e^{-j2v_2l_1/a}} - s_{11}}\\ &+ \frac{s_{12}^2\left(-2j\frac{l_1}{a}\frac{e^{+j2v_2l_1/a}}{\rho_{L1}} + \frac{\rho'_{L1}e^{+j2v_2l_1/a}}{\rho_{L1}^2} + s'_{11}\right)}{\left(\frac{1}{\rho_{L1}e^{-j2v_2l_1/a}} - s_{11}\right)^2}.\end{aligned} \tag{15}$$

The best situation occurs when $s'_{11} = s'_{12} = s'_{22} = 0$, corresponding to a junction whose parameters do not depend on frequency. Moreover, considering that $\rho_{L1}(v) \approx e^{-j\phi(v)}$ outside its band-pass and that $\phi'(v) \geq 0$ owing to the Foster's reactance theorem, by virtue of (14), we recover

$$\left|s_{12}^2\frac{\phi'(v) + 2\frac{l_1}{a}}{\left(\frac{1}{\rho_{L1}e^{-j2v_2l_1/a}} - s_{11}\right)^2}\right|_{v=v_2} \leq \frac{\varepsilon}{\Delta v} \tag{16}$$

which, in consideration of (2), reduces to

$$\left|s_{22}^2\frac{\phi'(v) + 2\frac{l_1}{a}}{s_{12}^2}\right|_{v=v_2} \leq \frac{\varepsilon}{\Delta v} \tag{17}$$

finally, recalling (9c) and (10), we obtain the following inequality

$$\left|\frac{s_{22}^2}{s_{12}^2}\right| = \left|\frac{\Gamma_e}{\frac{\Gamma_e - \Gamma_o}{2}}\right|^2 \leq \frac{\varepsilon}{\Delta v(\phi'(v) + 2l_1/a)} \quad v = v_2. \tag{18}$$

For an E-plane junction, we set now

$$\Gamma_e = |\Gamma_e|e^{j\phi_e} \tag{19a}$$

$$\Gamma_o = e^{j\phi_o}. \tag{19b}$$

For a H-plane one we set instead $\Gamma_e = e^{j\phi_e}, \Gamma_o = e^{j\phi_o}|\Gamma_o|$ and proceed along similar lines.

By defining $\Delta\phi = \phi_e - \phi_o$ and by substituting expressions (19) into (10), we obtain

$$\cos\Delta\phi = \frac{3|\Gamma_e|^2 - 1}{2|\Gamma_e|}. \tag{20}$$

The latter equation permits us to solve (18) with respect to the normalized bandwidth $2\Delta v$

$$2\Delta v \leq \frac{1 - |\Gamma_e|^2}{|\Gamma_e|^2}\frac{\varepsilon}{\phi'(v_2) + 2l_1/a}. \tag{21}$$

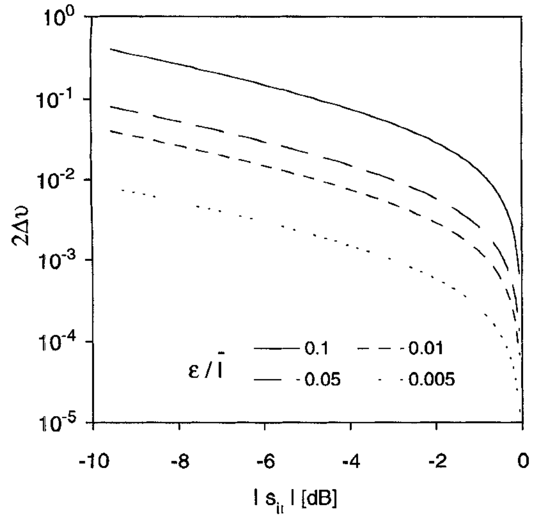


Fig. 4. Normalized bandwidth Δv versus the reflection coefficient at the common port of the three-port junction.

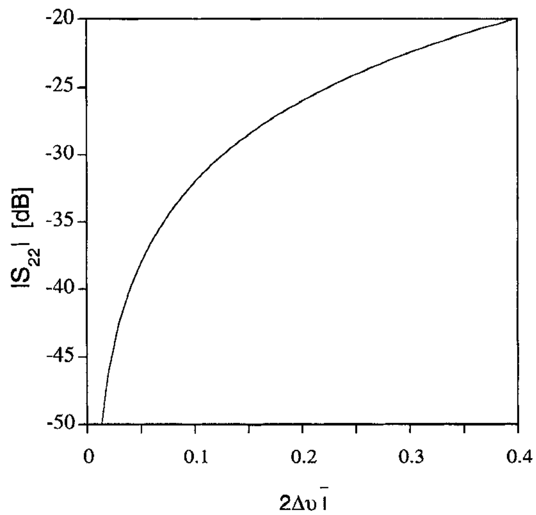


Fig. 5. Reflection coefficient at port 2 of the best junction possible ($|\Gamma_e| = 1/3$) when port 1 is closed on $F1$ (Fig. 2) versus $2\Delta v\bar{l}$.

Now, by introducing the effective normalized distance $\bar{l} = \phi'(v_2)/2 + l_1/a$ and remembering (9c), the theorem is proved. □

Inequality (21) gives the maximum normalized bandwidth, centered at f_2, of the two-port junction obtained by closing port 1 by filter $F1$ (Fig. 2), versus the reflection coefficient of the junction, as illustrated in Fig. 4. The ratio between ε, the maximum reflection coefficient admissible of the two-port so obtained, and the length $\bar{l}$, is taken as a parameter.

As we will see in detail in the Section III, the phase differential $\phi'(v_2)$ of the reflection of the load filter $F1$ plays a very important role in determining the maximum diplexer bandwidth; this variation being strongest near the passband for typical minimum phase filters, it is much more difficult to realize contiguous diplexers than noncontiguous ones.

The best situation occurs when $|\Gamma_e| = 1/3$. In that case, inequality (21) becomes

$$2\Delta v \leq \frac{4\varepsilon}{\bar{l}}. \tag{22}$$

On the contrary, it is noted that Δv tends to 0 as $|\Gamma_e|$ tends to 1.

The magnitude of the reflection for the two-port just described as a function of the bandwidth $2\Delta v\bar{l}$, as derived from (22), is shown in Fig. 5.

It is also interesting to note that in the most favorable case, the scattering matrix of the three-port junction assumes the following expression

$$\frac{1}{3}\begin{bmatrix} e^{j\phi} & -2e^{j\phi} & 2e^{j\xi} \\ -2e^{j\phi} & e^{j\phi} & 2e^{j\xi} \\ 2e^{j\xi} & 2e^{j\xi} & e^{j(2\xi-\phi)} \end{bmatrix}$$

ϕ, ξ being two constant phases. Finally note that the latter is the scattering matrix of an ideal Y-junction, which, therefore, seems to be the best choice with a view to designing diplexers with the criteria illustrated.

In conclusion, returning to the general case, the necessary conditions on the parameters of an optimum three-port symmetrical junction in order to obtain optimum perfomance when the junction is used in combination with two given filters in a diplexer configuration can be summarized by the following formulas

$$|s_{11}| = |s_{22}| \cong |s_{33}| \quad f = f_1, \ f = f_2 \tag{23a}$$

$$|s_{33}| \geq 1/3, \quad \text{preferably } |s_{33}| = 1/3 \tag{23b}$$

$$s'_{11} \cong s'_{12} \cong s'_{22} \cong 0 \quad f = f_1, \ f = f_2. \tag{23c}$$

The maximum achievable bandwidth of the diplexer is then stated by (21), and, moreover, having selected a junction with the above characteristics, the two filters have to be positioned as indicated by (6).

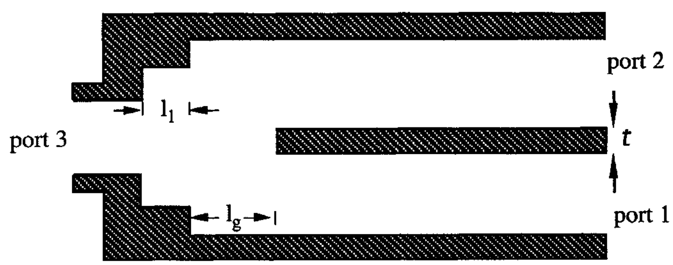


Fig. 6. E-plane section of the three-port junction employed in [1, 2].

III. Examples

In the following, we will demonstrate the validity of the foregoing criteria by means of two examples. The first one concerns a diplexer configuration designed, built and tested in [2], employing a symmetrical junction whose E-plane section is shown in Fig. 6.

This example was recomputed by making explicit use of the condition (5) within a simple optimization routine for the junction: The dimension we obtained are virtually identical to those reported in [2], as shown in Table I.

The results of the diplexer simulations are also virtually identical to those of Fig. 8 of [2]. Fig. 7 shows a comparison of the reflection magnitudes at ports 3 and 1 of the junction

TABLE I
COMPARISON BETWEEN DISTANCES CALCULATED IN [2] WITH THOSE CALCULATED BY OURSELVES BY OPTIMIZATION OF THE JUNCTION AND FORMULA (5)

distances	[2] (mm)	present method (mm)
l_1	2.434	2.430
l_g	12.977	12.975
c_1	9.074	9.066
c_2	8.285	8.280

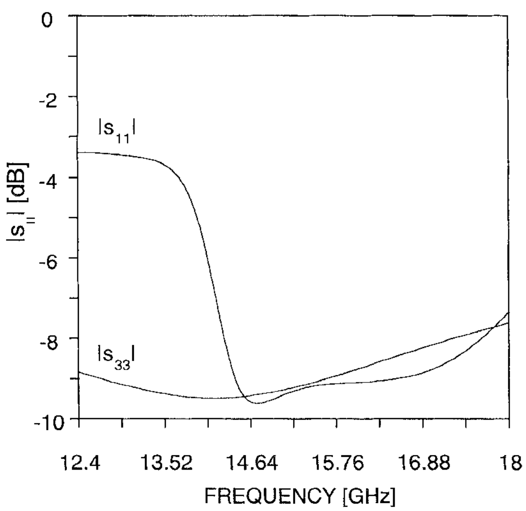


Fig. 7. Magnitudes of the reflection coefficients of the junction used in [2] at ports 1 and 3.

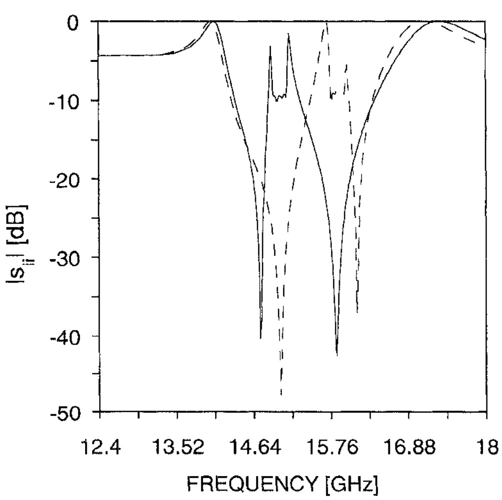


Fig. 8. Magnitude of the reflection coefficients for the junction of [2] at the common port 3 when port 1 is loaded by $F1$ (continuous) and port 2 is loaded by $F2$ (dashed).

without the filters; as it can be inferred, these values are very close to each other over the band of operation of the diplexer, as required by conditions (7), and they also satisfy

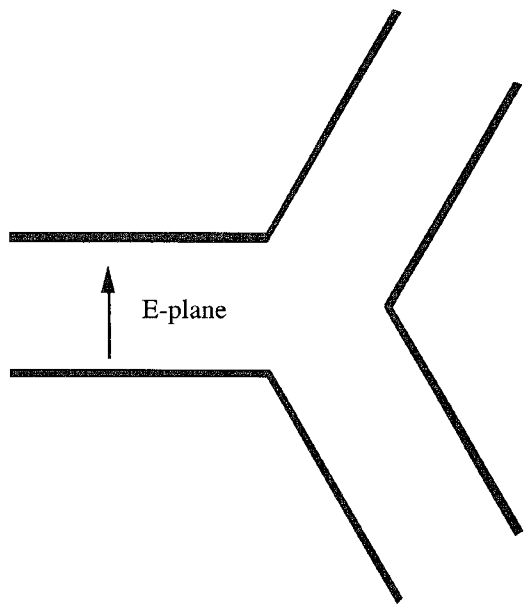


Fig. 9. E-plane section of the Y-junction.

conditions (23). Moreover, by application of (5a)–(5b), we recover exactly the same filter locations as were determined through numerical optimization by the authors of [2].

With reference to the same example, Fig. 8 compares reflection magnitudes appearing at port 3 when port 1 is loaded by $F1$ and port 2 matched (continuous) and when port 1 is matched and port 2 loaded by $F2$ (dashed). In the first case (continuous line) it is apparent a low-reflection window, centred in the midband frequency of the filter $F2$. Conversely, in the second case (dashed line), there is an analogous one about the passband of the filter $F1$. If the bandwidths of each filter exceed the widths of the windows (say at $-25 \div 30$ dB, depending on the diplexer specifications), a deterioration will occur. In that case, it is necessary either to redesign the junction or to optimize the filters.

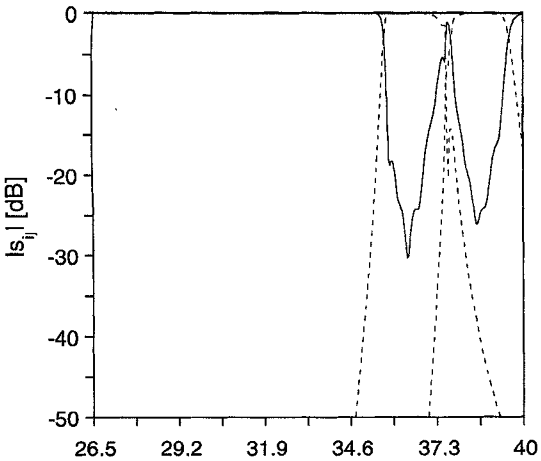
The second example consists of a simulation; we designed different Ka-band diplexers employing the same E-plane waveguide Y-junction (Fig. 9) modeled by the equivalent circuit of [11] and using given filters. The examples refer to wideband filters, this case being considerably more complicated than the narrow-band one [6].

We checked the method by considering first contiguous filters, then by increasing the separation between the bands of the two filters.

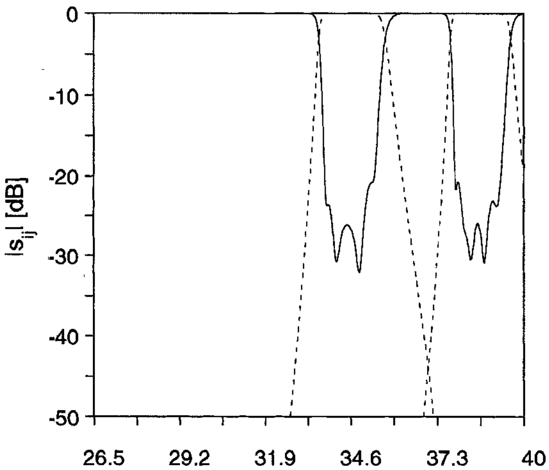
We start by evaluating the maximum bandwidth for which the two-port junction of Fig. 2 has a return loss larger than 26 dB, by considering the phase differentials of the filters to be negligible out of band. By substituting the magnitude of the reflection coefficient of the Y-junction, about 0.35, in (21) we obtain

$$2\Delta v \leq \frac{1-(0.35)^2}{(0.35)^2}\frac{0.05a}{2l}.$$

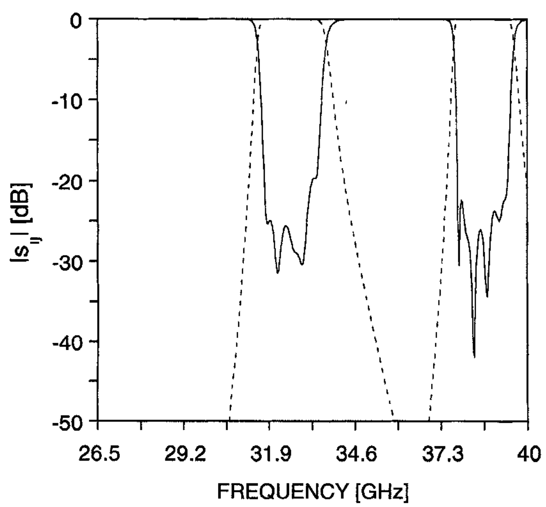
Supposing that $l \approx 4$ mm ($\approx \lambda_g/2$), we obtain $2\Delta\beta \leq 44.75$, corresponding to a bandwidth BW ≤ 1.79 GHz.



(a)



(b)



(c)

Fig. 10. Magnitude of the scattering parameters of the three simulated diplexers employing a Y-junction. Continuous lines: $|s_{33}|$, dashed lines: $|s_{13}|$ and $|s_{23}|$. (a) Filters $F1$ and $F2$, (b) filters $F1$ and $F3$, and (c) filters $F1$ and $F4$.

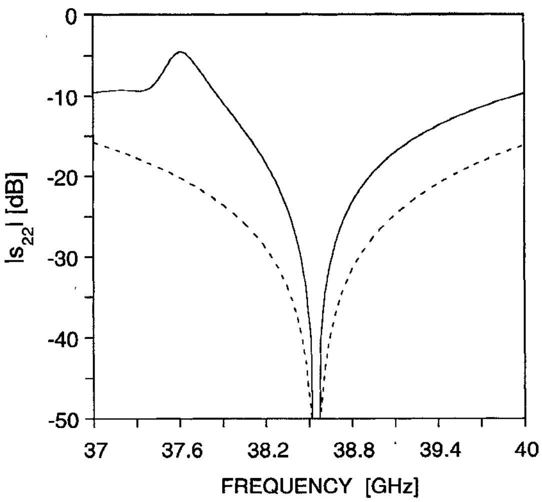


Fig. 11. Magnitude of the reflection coefficient at port 2 of the$Y$-junction when port 1 is closed on $F2$ (continuous) and $F4$ (dashed) respectively.

Having separately designed four E-plane septate 6 poles Tchebysheff filters $F1, \ldots, F4$ of 26 dB mrl and bandwidth 1.55 GHz, centered respectively at $f_1 = 38.5$ GHz, $f_2 = 36.5$ GHz, $f_3 = 34.5$ GHz, $f_4 = 32.5$ GHz, we combined filters and junction so as to obtain the three diplexers whose responses are shown in Fig. 10.

Note that we fixed the upper band filter and changed the lower one. In the first case (filters $F1$ and $F2$), the two filters are very close: There is an evident deterioration of the diplexer response compared with those of the filters that is due mainly to the excessive variation of the phase of one filter in the passband of the other. By measuring the slope of the phase of filter $F1$ at the frequencies f_2 and f_4, it is possible to give a more accurate estimate of the maximum bandwidth of the diplexer. In particular, in a neighborhood of f_2 and of f_4 we have, respectively, $\phi'(v(f_2)) \approx 3$ and $\phi'(v(f_4)) \approx 0.38$. corresponding to 480 MHz and 1.3 GHz bandwidths.

In the next example, we space further out the filter bandwidths. Figs. 10(b) and (c) refer to the diplexer obtained by connecting filters $F1$ and $F3, F1$ and $F4$, respectively, to the Y-junction: in both cases the solution is quite good and no further optimization is required.

Finally, we show the magnitudes of the reflection at port 2, when port 1 is loaded by $F2$ and $F4$, respectively, in a neighborhood of the midband frequency of $F1$ (Fig. 11): it is clear from the figure that the matching bandwidth depends on the phase behavior of the filter. We can also see that the bandwidth of the junction corresponding to a return loss of 26 dB is about 1.1 GHz when port 1 is loaded by $F4$ and just 330 MHz when port 1 is closed on $F2$, close enough to the value calculated above.

IV. Conclusion

We present simple analytical criteria for the design of optimum junctions for the realization of diplexers with given filters: these are of general validity, deriving solely from the properties of the scattering matrix of reciprocal, lossless symmetric three-port junctions.

The above criteria also provide explicitly two additional important parameters of the diplexer geometry, namely, the locations of the filters, resulting in a considerable simplification of the overall synthesis.

Results obtained by the application of the foregoing method to the synthesis are good except for the case in which the phase slope of one filter in the passband of the other is too large. Even in this unfavorable case the result can nonetheless be considered as a excellent starting point for an optimization routine.

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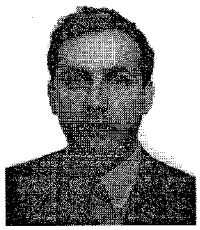

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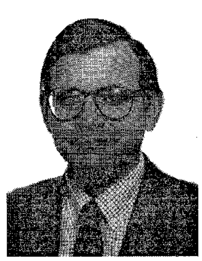

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