Microstrip-Slot Coupler Design—Part I: S-Parameters of Uncompensated and Compensated Couplers

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Abstract —**Using the even–odd mode analysis of four-port networks with double symmetry, the scattering parameters of the microstrip-slot coupler are derived from the even- and odd-mode parameters of the coupler cross section. The uncompensated coupler and the coupler compensated by extending the slot lines are treated and design specifications are given covering the compensation slot lines. A comparison with experimental data will be given in Part II.**

I. Introduction

THE MICROSTRIP-SLOT coupler in Fig. 1 is particularly suitable for the realization of 3-dB couplers in MIC technology. This coupler was originally proposed by de Ronde [1], and an empirical design was described by Garcia [2]. In addition, an analysis has been made by Schiek [3] with the aid of the equivalent circuit of the hybrid branchline coupler. This paper derives the scattering parameters of the microstrip-slot coupler from the even- and odd-mode parameters of the coupling section. The analysis is extended to include a lengthening of the slot line, which is used to compensate for the different phase velocities of the even and odd modes [4]. Design specifications for the compensated coupler are reported, and a comparison with experimental data is given in Part II.

II. Even–Odd Mode Analysis of the Coupler Without Compensation ($l_S = 0$)

The coupler of Fig. 1 consists of a coupling section C, connected by four feed lines of width w_p. The coupling section itself consists of a strip conductor under which a slot is symmetrically arranged (Fig. 1(c)), and reference planes T_1 and T_2 define the effective length l of this section with respect to the electrical behavior of the coupler. The slot line of length $l+2l_S$ is terminated in a disk-shaped open circuit of diameter D formed by removal of the ground plane metalization. When the position of T_1 and T_2, i.e., the distance d, is chosen appropriately, the effect of the parasitics due to the field perturbation occurring at the junctions at each end can be canceled out. For the following calculations, ideal junctions without parasitics are assumed at T_1 and T_2. Since the locations of these reference planes with respect to the ends of the strip conductor, i.e., d in Fig. 1(a), are not given by the theory, an experimental determination of the appropriate choice of d is given in Part II by comparing theoretical and measured electrical parameters of the coupler. The conductors are assumed to be lossless, and the open circuits at the ends of the slot line are assumed to be ideal. Also, the coupler without compensation is defined by $l_S = 0$. In this case, an ideal open

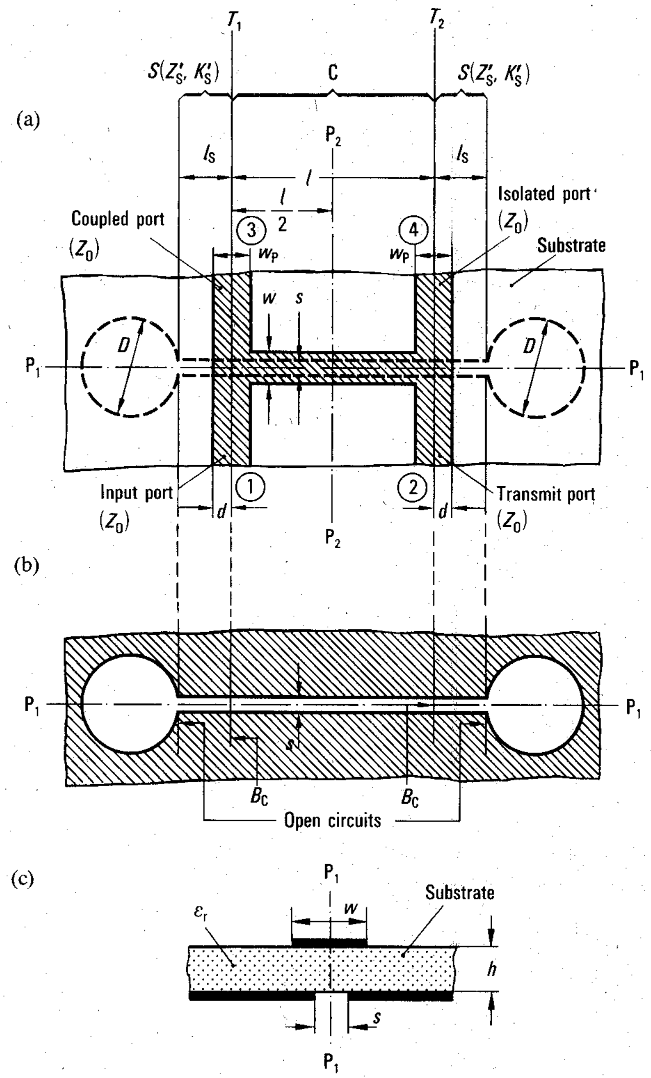

Fig. 1. Configuration of microstrip-slot coupler. (a) Upper side of substrate. (b) Bottom side of substrate. (c) Cross section.

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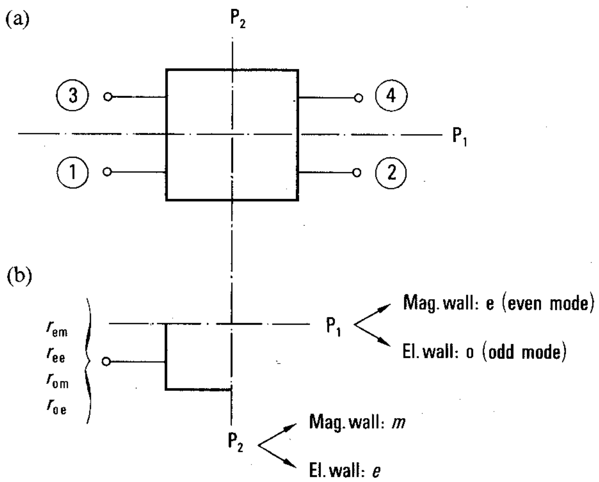

Fig. 2. Common four-port network with double symmetry with respect to P_1, P_2. (a) Configuration. (b) Definition of reflection coefficients.

circuit is located at the reference planes T_1 and T_2.

Note that this coupler configuration represents a reciprocal passive, linear four-port network with double symmetry with reference to the two symmetry planes P_1 and P_2 (Fig. 2(a)). Let Z_0 be the characteristic impedance of the connecting lines. The scattering parameters S_{ij} of the four-port network can be computed by an even–odd mode analysis [5], which is expanded to four-port networks in [6] and for the standard microstrip coupler in [7] and [8]. This yields for the present case

$$S_{11} = (r_{em} + r_{ee} + r_{om} + r_{oe})/4 \tag{1}$$

$$S_{21} = (r_{em} - r_{ee} + r_{om} - r_{oe})/4 \tag{2}$$

$$S_{31} = (r_{em} + r_{ee} - r_{om} - r_{oe})/4 \tag{3}$$

$$S_{41} = (r_{em} - r_{ee} - r_{om} + r_{oe})/4. \tag{4}$$

The parameters r_{em}, r_{ee}, r_{om}, and r_{oe} represent the input reflection coefficients, referenced to Z_0 appearing at each of the four terminals when certain combinations of magnetic and electric walls (Fig. 2(b)) are applied to the symmetry planes P_1 and P_2. The first subscript $e(o)$ denotes a magnetic (electric) wall at P_1, while the second subscript $m(e)$ denotes a magnetic wall or open circuit (electric wall or short circuit) at P_2.

To compute the reflection coefficients r_{em}, r_{ee}, r_{om}, and r_{oe} from the original coupler configuration in Fig. 1, magnetic or electric walls likewise have to be applied to the symmetry planes P_1, P_2. The conductor pattern of the coupling section is fed at its ends from the terminals 1, 3 or 2, 4 of the coupler four-port network as shown in Fig. 3(a). A magnetic wall at P_1 corresponds to even-mode excitation with equal terminal voltages $U_1 = U_3 = U$. P_1 then divides the conductor pattern into two identical transmission lines, each having the characteristic impedance Z_e and the effective permittivity $K_e = (c/v_e)^2$, where c denotes the light velocity and v_e the phase velocity of the even mode. In the case of the even-mode excitation, a quasi-microstrip mode [9]–[13] has a field pattern as shown in Fig. 3(b), a characteristic impedance Z_M, and an effective permittivity K_M. The relations are

$$Z_e = 2Z_M, \quad K_e = K_M. \tag{5}$$

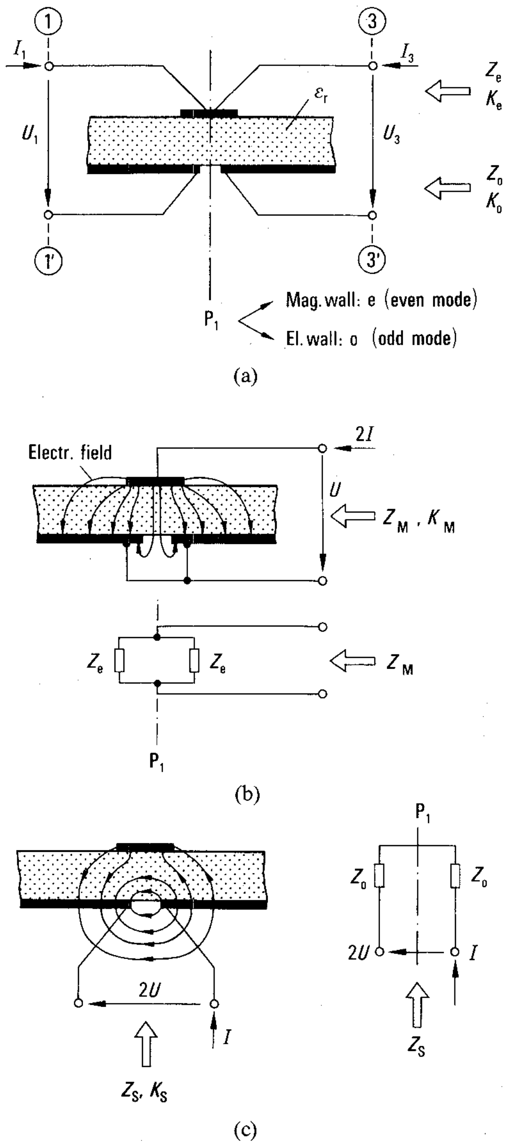

Fig. 3. Even–odd-mode excitation of coupling section. (a) Connection of terminals 1, 3 to coupling section. (b) Even–mode excitation with $U_1 = U_3 = U$, definition of Z_M, K_M (microstrip mode). (c) Odd-mode excitation with $U_1 = -U_3 = U$, definition of Z_S, K_S (slot-line mode).

An electric wall at P_1 corresponds to odd-mode excitation with opposite terminal voltages $U_1 = -U_3 = U$. The wall divides the conductor pattern of the coupling section into two transmission lines, each having the characteristic impedance Z_o and the effective permittivity $K_o = (c/v_o)^2$ (v_o denotes the phase velocity of the odd mode). A quasi-slot-line mode (quasi-TE-mode) now appears along the entire line with the characteristic impedance Z_S and the effective permittivity K_S (Fig. 3(c)) as analyzed in [10], [12], and [13]. The relations are

$$Z_o = \frac{Z_S}{2}, \quad K_o = K_S. \tag{6}$$

The reflection coefficients r_{em}, r_{ee}, r_{om}, and r_{oe} are the input reflection coefficients, referenced to the characteristic terminal impedance Z_0, of one of the four identical transmission lines in all four modes of excitation: "em", "ee", "om", and "oe". These transmission lines are cut out of the overall coupling section by the planes P_1, P_2. For $l_S = 0$ (coupler without compensation), these calculate at

$$r_{em} = \exp\left\{-j2\arctan\left[\frac{Z_0}{2Z_M}\tan\left(\frac{\Theta_e}{2}\right)\right]\right\} \tag{7}$$

$$r_{ee} = \exp\left\{j2\arctan\left[\frac{Z_0}{2Z_M}\cot\left(\frac{\Theta_e}{2}\right)\right]\right\} \tag{8}$$

$$r_{om} = \exp\left\{-j2\arctan\left[\frac{2Z_0}{Z_S}\tan\left(\frac{\Theta_o}{2}\right)\right]\right\} \tag{9}$$

$$r_{oe} = \exp\left\{j2\arctan\left[\frac{2Z_0}{Z_S}\cot\left(\frac{\Theta_o}{2}\right)\right]\right\} \tag{10}$$

with the electrical lengths $\Theta_e = \omega l\sqrt{K_M}/c$ and $\Theta_o = \omega l\sqrt{K_S}/c$. The transmission parameters S_{ij} of the uncompensated coupling section are in this way described as a function of Z_M, K_M, Z_S, K_S, length l, and the angular frequency ω by (1) through (10). These equations apply for arbitrary values of Z_0, Z_M, Z_S, and arbitrary phase velocities of v_e and v_o.

III. Even–Odd Mode Analysis of the Coupler with Compensation Lines ($l_S > 0$)

Now consider the coupling section C (Fig. 1), which includes the slot compensation length l_S according to [4]. Let S have the characteristic impedance Z_S', the effective permittivity K_S', and the length l_S. Z_S' and K_S' differ from the parameters Z_S and K_S of the coupling section with odd-mode excitation because, in the case of the added slot length l_S, the strip conductor is missing from the other side of the substrate (except for the section d if $d > 0$—this however will be neglected here). With typical couplers on alumina substrate ($Z_0 = 50\ \Omega, \epsilon_r = 9.8$), $K_S < K_M$ in all cases, whence $v_o > v_e$. For $l_S \ll \lambda/4$, the slot length acts as a shunt capacitance C_C (Fig. 4) that is only effective with odd-mode excitation. This capacitance increases the transmission phase between the terminals 1, 3 and 2, 4 for odd-mode excitation from φ_o to φ_o^*. For the selectable compensation frequency f_{co}, compensation can be realized with equal values of φ_o^* and the transmission phase for even-mode excitation $\varphi_e = \Theta_e = \omega l\sqrt{K_M}/c$ so that the directivity $D(f_{co}) = -20\log|S_{41}/S_{31}| = \infty$ and $S_{11}(f_{co}) = 0$. For even-mode excitation, $B_C = \omega C_C$ is ineffective.

Analysis is made by the even–odd mode method treated in Section II, whereby it is assumed, as shown in Fig. 4, that a susceptance

$$B_C = \omega C_C = \frac{1}{Z_S'}\tan\Theta_S' \tag{11}$$

with $\Theta_S' = \omega l_S\sqrt{K_S'}/c$, is shunted across the ends of the coupling section. Again, assume an ideal open circuit at the ends of the slot and an ideal junction at T_1, T_2. By applying appropriate magnetic and electric walls to the symmetry planes P_1, P_2, the reflection coefficients, referenced to Z_0, for odd-mode excitation can be computed from the input susceptance of the transmission lines shunted with the

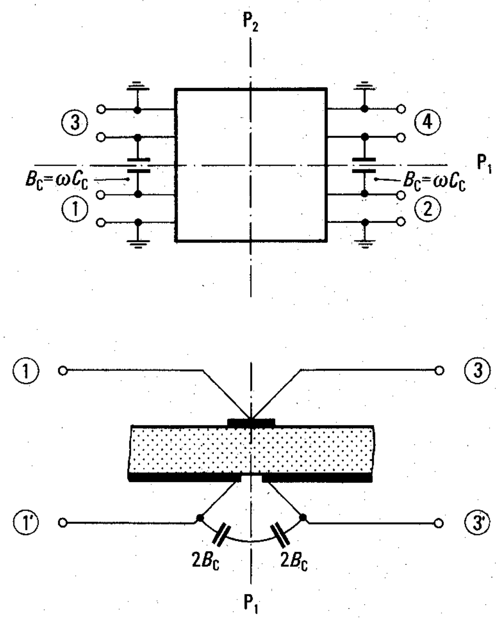

Fig. 4. Coupler with compensation lines $S(l_S > 0)$ represented by the input susceptance jB_C.

susceptance $2B_c$ at

$$r_{om} = \exp\left\{-j2\arctan\left[\frac{2Z_0}{Z_S}\tan\left(\frac{\Theta_o}{2}\right) + 2B_C Z_0\right]\right\} \tag{12}$$

$$r_{oe} = \exp\left\{-j2\arctan\left[-\frac{2Z_0}{Z_S}\cot\left(\frac{\Theta_o}{2}\right) + 2B_C Z_0\right]\right\} \tag{13}$$

with $\Theta_o = \omega l\sqrt{K_S}/c$ and B_C according to (11). The parameters r_{em} and r_{ee} can be computed with (7), and (8) because B_C is here ineffective. The scattering parameters S_{ij} of the coupler with $l_S > 0$ are computed with (1) through (4). These equations apply for arbitrary values of Z_0, Z_M, Z_S, K_M, and K_S, and arbitrary parameters Z_S', K_S', l_S.

IV. Special Cases of Microstrip-slot Couplers

A. Ideal Microstrip-Slot Coupler

The ideal microstrip-slot coupler cannot be realized with the real coupler configuration shown in Fig. 1 because in this case $K_S < K_M$. It can, however, be used to derive design equations for the real coupler. The ideal coupler is characterized by the reflection coefficient S_{11} being zero and S_{41} being zero (i.e., infinite directivity D) at all frequencies. Applying these conditions to (1) and (4) results in $r_{ee} = -r_{om}$ and $r_{em} = -r_{oe}$ for all frequencies. These equations apply only if two conditions with respect to the coupling section parameters exist. Condition 1 is that the coupling section has the same electrical length for the even and odd modes at all frequencies, i.e., $\Theta_e(f) = \Theta_o(f) = \Theta(f)$. This results in $K_e = K_o = K$ and $K_M = K_S = K$. Condition 2 is the matching condition

$$Z_0 = \sqrt{Z_e Z_o} = \sqrt{Z_M Z_S}. \tag{14}$$

With these conditions satisfied, (2), (3), and (7)–(10) for the transmission coefficient S_{21} and the coupling coefficient S_{31} assume the following simplified form:

$$S_{21}(f)=\frac{\sqrt{1-k^2}}{\sqrt{1-k^2}\cos\Theta+j\sin\Theta} \tag{15}$$

$$S_{31}(f)=\frac{jk\sin\Theta}{\sqrt{1-k^2}\cos\Theta+j\sin\Theta} \tag{16}$$

with $\Theta=\omega l\sqrt{K}/c$. These are the parameters of an ideal TEM coupler [14], where

$$k=\frac{Z_e-Z_o}{Z_e+Z_o}=\frac{4Z_M-Z_S}{4Z_M+Z_S} \tag{17}$$

is the coupling coefficient. At the center frequency $f_c=c/(4l\sqrt{K})$, $l=\lambda/4$, where λ is the line wavelength. The magnitude of $S_{21}(f_c)=-j\sqrt{1-k^2}$ is here at minimum and that of $S_{31}(f_c)=k$ is at maximum. For a given center-frequency coupling loss $a_c=-20\log k$ and a given characteristic impedance Z_0, (14) and (17) yield the synthesis equations

$$Z_M=\frac{Z_0}{2}\sqrt{\frac{1+k}{1-k}} \tag{18}$$

$$Z_S=2Z_0\sqrt{\frac{1-k}{1+k}}=\frac{Z_0^2}{Z_M}. \tag{19}$$

For a 3-dB coupler with $Z_0=50\ \Omega$, the characteristic impedances $Z_M=60.35\ \Omega$ and $Z_S=41.4\ \Omega$ are obtained.

At this point it is important to consider the differences between the present analysis of the microstrip-slot coupler and the hybrid branchline coupler analysis [3]. In the latter, the microstrip-slot coupler is treated as a special case of the π-type hybrid branchline coupler as shown in Fig. 5(a). This consists of two identical parallel lines G_p and a series transmission line G_s, each having the length l. To allow comparison of the two analyses, the hybrid branchline coupler is subjected to even-mode excitation (magnetic wall at P_1) and odd-mode excitation (electric wall at P_1) according to Fig. 5(b) for identical phase velocities in all lines. It follows from this that $Z_e=Z_p$ for even-mode excitation and Z_o is represented by Z_p in parallel to $Z_{se}/2$, resulting in

$$Z_o=\frac{Z_pZ_{se}}{2Z_p+Z_{se}} \tag{20}$$

for odd-mode excitation. This results in the synthesis equations $Z_p=Z_0\sqrt{(1+k)/(1-k)}$ and $Z_{se}=Z_0\sqrt{1-k^2}/k$ for the hybrid branchline coupler, which yield $Z_p=120.7\ \Omega$ and $Z_{se}=50\ \Omega$ for a 3-dB coupler with a Z_0 of 50 Ω. The analysis of this paper and that of Schiek's paper [3] agree if the microstrip-mode-excited coupling section of Fig. 3(b) with the characteristic impedance Z_M is modeled by a parallel combination of two impedances Z_p, and the slot-mode-excited coupling section of Fig. 3(c) with the characteristic impedance Z_S is modeled by a parallel combination of Z_{se} and $2Z_p$. Applied to the above mentioned 3-dB coupler with $Z_0=50\ \Omega$, a parallel combination of two impedances $Z_p=120.7\ \Omega$ results in an impedance of 60.35 Ω, which corresponds to Z_M according to (18), and a

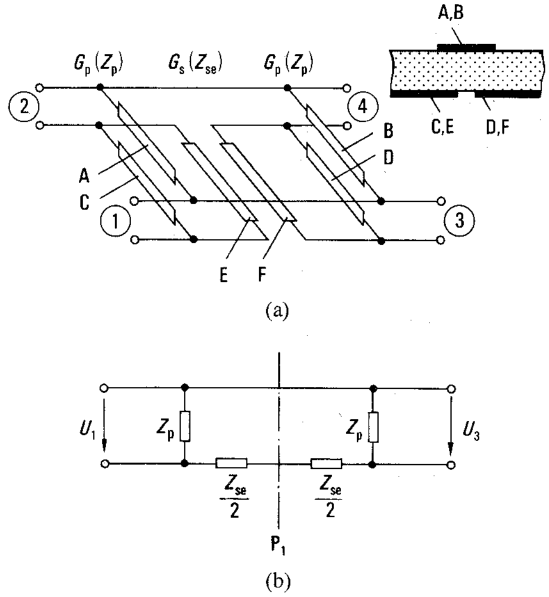

Fig. 5. Comparison with hybrid branchline coupler analysis according to [3]. (a) Equivalent diagram of hybrid branchline coupler. (b) Even–odd mode excitation of (a).

parallel combination of $Z_{se}=50\ \Omega$ and $2Z_p=241.4\ \Omega$ results in an impedance of 41.4 Ω, which corresponds to Z_S according to (19).

A general rigorous analysis of the coupling section, such as a numerical analysis from the field quantities permits—basically—only the calculation of the even- and odd-mode impedances Z_e and Z_o. Since in Schiek's equivalent circuit Z_o is modeled by a combination of both Z_p and Z_{se}, it is not possible to calculate Z_{se} separately from the odd-mode-excited coupling section, i.e., from the field quantities. Therefore, the interpretation of Z_{se} as the isolated slot line impedance, which has been adopted in Section IV-1 of [3], where a 50-Ω slot-line impedance is given for the 3-dB coupler with $Z_0=50\ \Omega$, is strictly speaking an approximation. However, the characteristic impedances Z_M and Z_S of the analysis of this paper are defined directly from the field quantities and therefore characterize the coupling section exactly.

B. Real Uncompensated Coupler

For the uncompensated coupler, $l_S=0$ (Fig. 1). Assuming commercial couplers of conventional dimensions, $K_S<K_M$ in all cases, whence $v_o>v_e$, whereby, however, the difference $\Delta K/K_m\ll 1$, with $K_m=(K_M+K_S)/2$ and $\Delta K=K_M-K_S$. This means that, for feeding at terminal 1, the assignment of the transmission path to terminal 2, of the coupling path to terminal 3 and of the isolated port to terminal 4 remain unchanged. The coupler parameters are no longer ideal, especially its directivity, which is $D<\infty$ and $|S_{11}|>0$. However, $|S_{11}|,|S_{41}|\ll 1$ remain if the matching condition (14) is taken into account or the synthesis equations (18) and (19) of the ideal coupler are used for designing. The scattering parameters are computed with (1)–(4) and (7)–(10), whereby $S_{21}(f)$ and $S_{31}(f)$ closely approximate the ideal coupler parameters computed with (15) and (16). For the center frequency $f_c=$

$c/\left(4l\sqrt{K_m}\right)$ the reflection coefficient and the directivity can be obtained from (1)–(4), (7)–(10) by a first order approximation, resulting in

$$S_{11}(f_c) = -j\frac{\pi}{4}k\sqrt{1-k^2}\,\frac{K_M - K_S}{K_M + K_S} \tag{21}$$

$$D(f_c) = -20\log\left|\frac{\pi}{4}\left(\frac{1}{k} - k\right)\frac{K_M - K_S}{K_M + K_S}\right|. \tag{22}$$

The essential performance resembles that of the microstrip coupler [7].

C. Real Compensated Coupler

For real couplers with $K_S < K_M$, matching and compensation, i.e., $S_{11} = 0$, $S_{41} = 0$, can be realized with the aid of the compensation lines S only at a single arbitrarily chosen frequency, viz., the compensation frequency f_{co}. For this it is essential that 1) the matching condition (14) be satisfied, 2) Z_S' and l_S be appropriately chosen, and, 3) that Z_S be increased slightly to Z_S^*. To derive the compensation conditions, we refer to the section entitled "*Ideal Microstrip-Slot Coupler*", according to which ideal coupler behavior at f_{co} can be obtained if the matching condition (14) is satisfied and K_S is equal to K_M. For the real coupler, the matching condition can easily be satisfied by appropriate choice of the characteristic impedances Z_M and Z_S according to the synthesis equations (18) and (19), but still $K_S < K_M$ because of the different field distributions of the microstrip and slot modes. At one particular frequency, however, namely at the compensation frequency f_{co}, the effect of the differences in K_S and K_M can be compensated if we extend the slot line by a certain amount l_S (Fig. 1). The additional slot lines S resulting from this extension leave the transmission characteristics of the microstrip-mode-excited coupling section unchanged, because here the voltage across the slot of the coupling section is zero. These additional lines S modify the transmission characteristics of the slot-mode-excited (Fig. 3(c)) coupling section only. This real slot-mode-excited coupling section with compensation lines, hereinafter referred to as (R), is shown in Fig. 6(a). To obtain compensation at f_{co}, the transmission phase of (R) applying to $l_S = 0$ has to be increased by adding capacitive end loading. This is accomplished by lengthening the slot by a certain amount l_S, at which this transmission phase reaches the transmission phase $2\pi l f_{co}\sqrt{K_M}/c$ of the microstrip-mode-excited coupling section. Unfortunately, however, the capacitive end loading decreases the input impedance level of (R). To compensate for this, the characteristic impedance Z_S has to be increased to Z_S^*.

Now assume the characteristic impedance of lines S to be $Z_S' \cong Z_S^*$. To derive equations for l_S and Z_S^*, the S parameters $S_{aa}^{(R)}$ and $S_{ab}^{(R)}$ of (R) have, at f_{co}, to be respectively equated to the S parameters $S_{aa}^{(I)}$ and $S_{ab}^{(I)}$ of an ideal slot-mode-excited coupling section without compensation lines (Fig. 6(b)), hereinafter referred to as (I), with the parameters Z_S according to (19) and $K_S^{(I)} = K_M$. (I) would provide compensation at any frequency. Therefore, the compensation conditions are $S_{aa}^{(R)} = S_{aa}^{(I)}$ and $S_{ab}^{(R)} = S_{ab}^{(I)}$ at f_{co}, all S parameters being referred to Z_S. The terminals a and b are the ends T_1 and T_2 of the coupling section. Again, utilizing the symmetry of the configurations (R) and (I), we can calculate the S parameters S_{aa}, S_{ab} from one-port reflection coefficients r_m and r_e, as we did for the whole coupler in (1)–(4). The above mentioned compensation conditions can then be divided into two alternative compensation conditions

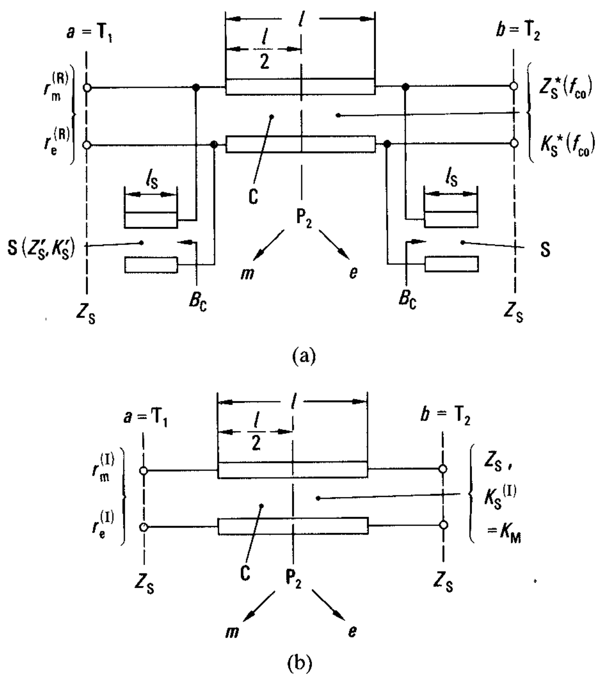

Fig. 6. For synthesis of compensation conditions. (a) Real coupling section (R) with compensation lines S excited with slot-line mode. (b) Fictitious ideal coupling section (I) with $K_S^{(I)} = K_M$ excited with slot line mode.

$$r_m^{(R)}(f_{co}) = r_m^{(I)}(f_{co}) \tag{23}$$

$$r_e^{(R)}(f_{co}) = r_e^{(I)}(f_{co}) \tag{24}$$

where $r_m^{(I)}$ and $r_e^{(I)}$ are the input reflection coefficients, referenced to Z_S, of the ideal configuration if a magnetic (subscript m) or electric (subscript e) wall is applied to P_2. The parameters $r_m^{(R)}$ and $r_e^{(R)}$ are the corresponding reflection coefficients, likewise referenced to Z_S, of the real configuration. Equations (23) and (24) can only be satisfied by altering the characteristic impedance of the slot line.

Now assume that compensation has to be realized at the center frequency referenced to the microstrip transmission line, i.e.,

$$f_{co} = f_M = \frac{c}{4l\sqrt{K_M}}. \tag{25}$$

Further assume, for simplicity, that the slight increase in Z_S to Z_S^* required for compensation will not alter the effective permittivity K_S, i.e., $K_S^* = K_S$. Thus (23) and (24) will yield the compensation conditions

$$-2\arctan\left[B_{C,co}\cdot Z_S - \frac{Z_S}{Z_S^*}\cot\left(\frac{\pi l f_M\sqrt{K_S}}{c}\right)\right] = \frac{\pi}{2} \tag{26}$$

$$2\arctan\left[B_{C,co}\cdot Z_S+\frac{Z_S}{Z_S^*}\tan\left(\frac{\pi l f_M\sqrt{K_S}}{c}\right)\right]=\frac{\pi}{2} \quad (27)$$

with the input susceptance (obtaining at the compensation frequency $f_{co}=f_M$)

$$B_{C,co}=2\pi f_M\cdot C_{C,co}=\frac{1}{Z_S^*}\tan\left(\frac{2\pi l_S f_M\sqrt{K_S}}{c}\right) \quad (28)$$

of the compensation lines S. In (28) it is assumed that the characteristic impedance of these lines is $Z_S'=Z_S^*$ and that their effective permittivity is $K_S'=K_S$.

Equations (26) and (27) yield, after rearrangement, the increased characteristic impedance required for compensation at the compensation frequency $f_{co}=f_M$

$$Z_S^*=\frac{Z_S}{2}\left[\cot\left(\frac{\pi}{4}\sqrt{\frac{K_S}{K_M}}\right)+\tan\left(\frac{\pi}{4}\sqrt{\frac{K_S}{K_M}}\right)\right]. \quad (29)$$

The length of the compensation lines, which are assumed to have approximately the parameters $Z_S'=Z_S^*$ and $K_S'=K_S^*=K_S$ should be chosen according to

$$l_S=l\sqrt{\frac{K_M}{K_S}}\,\frac{2}{\pi}\arctan\left\{\frac{Z_S^*}{Z_S}\left[1-2\sin^2\left(\frac{\pi}{4}\sqrt{\frac{K_S}{K_M}}\right)\right]\right\}. \quad (30)$$

As an alternative to extending the slot line by l_S it is possible to shunt both ends of the coupling section at T_1, T_2 with a capacitance

$$C_{C,co}=\epsilon_o\frac{\eta_o}{Z_S}\frac{2}{\pi}l\sqrt{K_M}\left[1-2\sin^2\left(\frac{\pi}{4}\sqrt{\frac{K_S}{K_M}}\right)\right] \quad (31)$$

parallel to the slot line ($\eta_o=120\pi\ \Omega$, $\epsilon_o=1/(3.6\pi)$ pF/cm). The values of l_S according to (30) agree to within a few percent with the first-order approximation given in [4]. [4], however, does not account for the increase in the characteristic impedance Z_S according to (29).

V. Conclusions

Using the even–odd mode analysis for four-port networks with double symmetry, equations have been derived for the frequency-dependent scattering parameters S_{ij} of microstrip-slot couplers, whereby different phase velocities v_e, v_o of the even and odd modes, as well as the mismatch, are included through the arbitrary choice of the characteristic impedance Z_0 and the even-mode and odd-mode characteristic impedances Z_e, Z_o. The analysis was extended to cover the coupler with supplementary slot lines as conventionally used for compensation. For $v_e=v_o$ the parameters of an ideal TEM coupler are realized. Simple design specifications are derived for the compensation slot lines.

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References

[1] F. C. de Ronde, "A new class of microstrip directional couplers," in *Proc. IEEE Int. Microwave Symp.*, 1970, pp. 184–186.
[2] J. A. Garcia, "A wide-band quadrature hybrid coupler," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 660–661, July 1971.
[3] B. Schiek, "Hybrid branchline couplers—A useful new class of directional couplers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 864–869, Oct. 1974.
[4] B. Schiek and J. Koehler, "Improving the isolation of 3-dB couplers in microstrip-slotline technique," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 5–7, Jan. 1978.
[5] J. Reed and G. J. Wheeler, "A method of analysis of symmetrical fourport networks," *IRE Trans. Microwave Theory Tech.*, vol. MTT-4, pp. 246–252, Oct. 1956.
[6] J. L. Altman, *Microwave Circuits.* Princeton, NJ: Van Nostrand, 1964, p. 115.
[7] H. E. Brenner, "Perturbations of the critical parameters of quarter-wave directional couplers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-15, pp. 384–385, June 1967.
[8] G. Schaller, "Directivity improvement of microstrip λ/4-directional couplers," *Arch. Elek. Übertragung.*, vol. 26, pp. 508–509, Nov. 1972.
[9] R. H. Jansen, "Microstrip lines with partially removed ground metallization, theory, and applications," *Arch. Elek. Übertragung.*, vol. 32, pp. 485–492, Dec. 1978.
[10] H. Ogawa and M. Aikawa, "Analysis of coupled microstrip-slot lines," *Trans. of the IECE Jap.*, vol. E 62, pp. 269–270, Apr. 1979.
[11] T. Itoh, D. Ratliff, and A. S. Hebert, "Generalized spectral domain method for multi-conductor printed lines and its application to tunable suspended microstrips," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1978, pp. 76–78.
[12] W. Schumacher, "Wellentypen und Feldverteilungen auf unsymmetrischen planaren Leitungen mit inhomogenem Dielektrikum," *Arch. Elek. Übertragung.*, vol. 34, pp. 445–453, Nov. 1980.
[13] K. Sachse *et. al.*, "Waveguide treatment of the suspended microstrip line with tuning septums using the spectral domain approach and the finite-element method," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1980, pp. 468–470.
[14] B. M. Oliver, "Directional electromagnetic couplers," in *Proc. IRE*, vol. 42, pp. 1686–1692, Nov. 1954.

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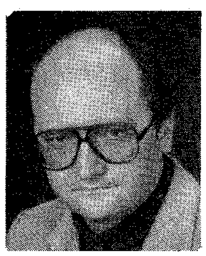

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