Accurate Analysis of Coupled Strip-Finline Structure for Phase Constant, Characteristic Impedance, Dielectric and Conductor Losses

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Abstract —Propagation constant, characteristic impedance, dielectric loss, and conductor loss of coupled strip-unilateral finline is here computed for the first time. The technique of analysis is based on the assumption of hybrid wave propagation implemented through the spectral domain approach for the phase constant. A perturbation method together with the spectral analysis has been applied to find the losses. The basis functions used to approximate fields within unilateral finline gap and currents on the strip have been selected as Legendre polynomials for the unbounded field or current and trigonometric functions for bounded field or current. The Green's function matrix in the spectral domain for the two distinct planes of the coupled strip-unilateral finline has also been presented. This gives the opportunity for direct implementation in the analysis of other similar structures. The possibility of the extension of the technique to shielded stratified dielectric with distributed planar conductor within different layers has been also discussed.

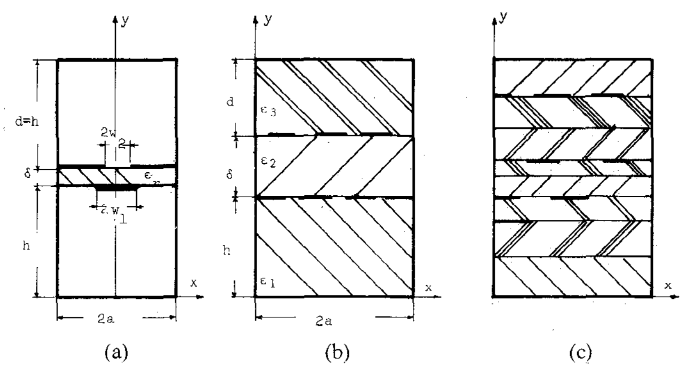

Fig. 1. (a) Coupled microstrip-unilateral finline. (b) Two-layer arbitrarily distributed strip structure. (c) Multilayer dielectric and strip structure.

I. INTRODUCTION

IN SPITE OF considerable usage of coupled strip-unilateral finline, Fig. 1(a), in microwave integrated circuits, particularly at millimeter band, no account of phase constant, characteristic impedance, or dielectric and conductor losses has been so far given for this structure. However, a few authors have solved the problem of coupled microstrip-slotline which has similarity to strip-finline structure [1]–[3], but in these works either the effect of sidewalls is neglected or the role of imperfection in dielectric and conductors on the field strength is ignored. In this paper, a rigorous hybrid mode solution based on the spectral domain technique is presented for strip-finline structure. From this an explicit form of $[G]$ matrix, similar to that given in [4] and used in [5] for other finline structures is presented. This matrix, due to the increase of one layer of conductor to its previous distribution shown in [4], has the order 4. Generally speaking, the increase of every one conductor layer corresponds to an increase of two to the order of $[G]$, Fig. 1(c). Therefore, a multi-dielectric layer structure with four different conductor layers would have a $[G]$ matrix of the order 8. It was found, out of 16 elements of $[G]$ matrix of structures with two-layer conductors embodied in three dielectric regions, Fig. 1(b), that only 9 of them are independent. Obviously, this matrix degenerates into a $[G]$ matrix of slot or strip or their suspended versions [6], depending upon the geometry of the configuration of the line. This point becomes valuable when the technique is to be checked against other well-established programs of microstrip and slot line.

As far as the basis functions for approximation of field and current are concerned, those given in [1], [3] for microstrip-slot lines may be used for the simplest solution: the zeroth-order solution. However, this choice is limited to those coupled strip-finlines whose strip and slot width are very small compared with the shield width. Furthermore, higher order modes cannot be predicted by the mentioned basis functions. A general form of basis functions satisfying the edge condition explicitly for unbounded fields and currents is presented in [2]. Unfortunately, this explicit treatment of the edges does not permit the normal perturbation analysis of conductor loss [5]. Therefore, to fulfill the requirements of conductor loss, as well as dispersion calculations, Legendre polynomials are again [5], [7] chosen to approximate the unbounded field or current. For the bounded field and current, trigonometric functions have been taken as basis functions. Losses for strip-finline structure are evaluated through a perturbation technique introduced in [5], [6], [7]. Accuracy of the results depends upon the low dielectric loss tangent and the high conductivity of metal.

The definitions of characteristic impedances adopted are

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current-power relation for the microstrip mode and voltage–power relation for the slot mode. These were found earlier [6] to be most appropriate for the quasi-TEM and quasi-waveguide modes, respectively. However, these definitions are not unique because of the mixed dielectric structure.

II. Theory

Consider Fig. 1(b) showing a shielded three-layer dielectric structure in which two planes of distributed conductors are arranged as desired. Therefore, coupled strip-finline structure, Fig. 1(a), emerges as a subclass of Fig. 1(b). For the sake of brevity, since the method of analysis assumes hybrid wave propagation through the structure, implemented via the spectral domain approach to obtain the phase constant, the general treatment presented by the authors in [4], [6], [7] is considered as the basis of analysis. In [4], [6] all the scalar potential functions, their transformed forms, and the boundary conditions are clearly stated. The only difference in the new configuration, Fig. 1(b), is the second layer of conductors. Following the same procedure as in [4], [6], after eliminating the unnecessary unknown parameters and retaining only $\tilde{E}_{x,h}$, $\tilde{E}_{z,h}$, $\tilde{E}_{x,d}$, $\tilde{E}_{z,d}$, $\tilde{J}_{x,h}$, $\tilde{J}_{z,h}$, $\tilde{J}_{x,d}$, and $\tilde{J}_{z,d}$ the $[G]$ matrix relating these fields and currents is given as

$$\left[\begin{array}{cc|cc} -G_{1,4} & -G_{1,3} & G_{1,2} & G_{1,1} \\ -G_{1,3} & -G_{2,3} & G_{2,2} & G_{1,2} \\ \hline G_{1,2} & G_{2,2} & G_{3,2} & G_{3,1} \\ G_{1,1} & G_{1,2} & G_{3,1} & -G_{4,1} \end{array}\right] \begin{bmatrix} \tilde{J}_{z,h} \\ \tilde{J}_{x,h} \\ \tilde{E}_{x,d} \\ \tilde{E}_{z,d} \end{bmatrix} = \begin{bmatrix} \tilde{E}_{z,h} \\ \tilde{E}_{x,h} \\ \tilde{J}_{x,d} \\ \tilde{J}_{z,d} \end{bmatrix} \tag{1}$$

Elements of $[G]$ have been introduced in the Appendix. As seen from $[G]$ in (1) only 9 elements are independent. In (1) $\tilde{J}_{z,h}$, $\tilde{J}_{x,h}$, $\tilde{E}_{z,h}$, and $\tilde{E}_{x,h}$ are currents and fields at $y=h$ while $\tilde{J}_{z,d}$, $\tilde{J}_{x,d}$, $\tilde{E}_{z,d}$, and $\tilde{E}_{x,d}$ are the same parameters at $y=h+\delta$. $\sim$ denotes that all the fields and currents are in the Fourier domain. As (1) is written for a two-layer conductor structure, it is reduced to simpler forms, like in [4], [6], if either $h=0$ or $d=0$ or the strip or fins disappear.

At this stage, it is worthwhile examining the possibility of the extension of the method to stratified dielectric with multilayer strip, Fig. 1(c). In fact by using the transfer matrix [4] it is always possible to relate the fields coefficients of one dielectric layer adjacent to one arbitrary strip layer to the nearest dielectric region next to the subsequent strip layer. Thus if the procedure is continued for all the layers of dielectric media and the boundary conditions at each conductor layers are also applied, the final result in terms of all the unknown fields and currents of all conductor interfaces can be represented through a matrix relation like (1). Clearly, the order of the associated $[G]$ matrix is two times the number of planes containing conducting strips.

Relation (1) can be converted into a linear system of homogenous equations by choosing first a set of basis functions for $\tilde{E}_{x,d}$, $\tilde{E}_{z,d}$, $\tilde{J}_{x,h}$, and $\tilde{J}_{z,h}$. This set for coupled strip-unilateral finline is given as follows:

$$J^e_{x,h} \text{ or } E^o_{z,d} = \sum_{m=1}^{M} a_m \sin(m\pi x/w), \qquad |x|<w' \tag{2}$$

$$J^e_{z,h} \text{ or } E^o_{x,d} = \sum_{n=1}^{N} b_n P_{2(n-1)}(x/w), \qquad |x|<w \tag{3}$$

$$E^e_{z,d} \text{ or } J^o_{x,h} = \sum_{p=1}^{P} c_p \cos(2p-1)x/w', \qquad |x|<w' \tag{4}$$

$$E^e_{x,d} \text{ or } J^o_{z,h} = \sum_{q=1}^{Q} d_q P_{2q-1}(x/w'), \qquad |x|<w'. \tag{5}$$

In the above relations, superscripts o and e represent odd and even modes of the coupled strip-unilateral finline structure. In (2)–(5), w and w' are given by

$$\left.\begin{array}{l} w=w_2 \\ w'=w_1 \end{array}\right\}, \quad \text{for the odd mode} \tag{6}$$

$$\left.\begin{array}{l} w=w_1 \\ w'=w_2 \end{array}\right\}, \quad \text{for the even mode.} \tag{7}$$

In the transformation of the basis function, one should be careful to observe that for the even mode (microstrip mode with a magnetic wall plane of symmetry), the spectrum parameter α_n is

$$\alpha_n = (n+1/2)\pi/a, \qquad n=0,1,2,\cdots \tag{8}$$

and the same parameter for the odd mode (finline mode), is given by [5]

$$\alpha_n = n\pi/a, \qquad n=0,1,2,\cdots. \tag{9}$$

Substituting the transformed version of (2)–(5) into (1) and applying Galerkin's method and Parseval's identity, the final set of linear homogeneous equations, very similar to those previously derived for microstrip and so on [6]–[9], results. The nontrivial solution of this system gives the phase constant and consequently all the fields in the Fourier or space domain.

A. Dielectric Loss

A perturbation method developed for small loss dielectrics in multilayer planar structures [6], [7] can be directly applied to the coupled strip finline structure or in general to any substructure of Fig. 1(c). The attenuation at frequency ω due to dielectric layers, each having a loss tangent of $\tan\delta_i$ and a cross section of S_i, can be given by [6], [7]

$$\alpha_d = \frac{\omega \sum_i \epsilon_i \tan\delta_i \iint_{S_i} |\vec{E}_0|^2\, dS}{2\,\mathrm{Re} \iint_S \vec{E}_0 \times \vec{H}_0^*\, d\vec{S}} \tag{10}$$

where

$$i,\ S,\ \vec{E}_0, \text{ and } \vec{H}_0$$

are the dielectric index, the whole cross section, and the unperturbed fields given by the spectral domain approach.

B. Conductor Loss

For attenuation due to conductor loss for good conductors, the following conventional formula derived by a perturbation technique is given [7]

$$\alpha_c = \frac{R_s \int_c |\vec{H}_t|^2 \, d\ell}{2 \operatorname{Re} \iint_S \vec{E}_0 \times \vec{H}_0^* \, d\vec{S}} \tag{11}$$

where R_s is the surface resistance and $\vec{H}_t$ is the magnetic tangent field around the conductors for the lossless case. However, in (11), near the edge of any infinitely thin conductor, $\vec{H}_t$ for the lossless case becomes unbounded and the integral ceases to exist. This situation occurs in the analysis of all planar structures of Fig. 1(a), (b), (c), with the assumption of negligible strip thickness [5]. This difficulty has been overcome by using bounded basis functions in (3) and (5), and the rationale is described in [5]. To achieve accuracy with (11), the following inequality should be considered

$$\text{skin depth} \ll \text{strip thickness} \ll \text{dielectric thickness.}$$

For further information, different comparative studies have been carried out between α_c from (11) and those analytically available in [5].

C. Characteristic Impedance

Since a unique and general definition of characteristic impedance for the structure of Fig. 1(c) is not available at high frequencies, for the sake of definiteness, the following two definitions are given for coupled strip-unilateral finline structure

$$Z_o = V^2/P \text{ for slot mode} \tag{12}$$

$$Z_e = P/I^2 \text{ for microstrip mode} \tag{13}$$

where V and I are integrations of E_x across the slot and J_z over the strip, respectively, and P is the transmitted power.

III. Numerical Results

The implementation of all the above expressions in a computer program gives the opportunity of fast and accurate analysis of coupled strip-unilateral finlines.

The first example treated by the program is, however, aimed to check whether $[G]$ in (1) properly degenerates into $[G]$ of microstrip or slot line. For this purpose, a shielded coupled strip-slot has been analyzed and the results of even and odd modes for different values of w_1 and w_2 have been shown in Fig. 2. As noticed in the figure, when w_1 approaches zero for the odd mode and w_2 approaches a for the even mode, the normalized phase constants approach phase constants of shielded slot and suspended microstrip lines, respectively. Results for shielded slot and shielded suspended microstrip were obtained from [6] and [10], respectively, where in turn they had been checked against [11].

In the second example, which is a coupled strip-unilateral finline built in WG-22 waveguide, Fig. 1(a), not only the normalized wavelengths for the even and odd modes are given but also the associated characteristic impedances and losses are computed. Furthermore, for the even mode solution, all the parameters obtained near the two limits

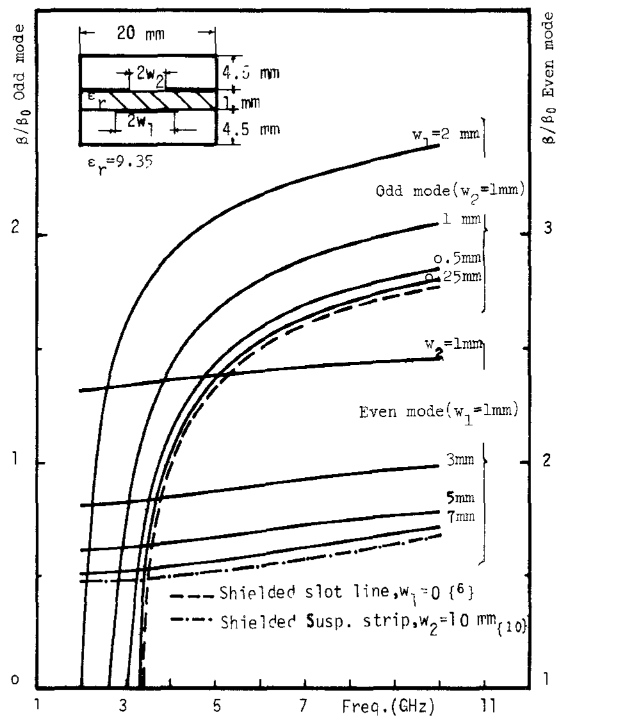

Fig. 2 Even and odd mode propagation characteristics of shielded microstrip-slot line and their asymptotic approaches to shielded suspended microstrip for the even mode and to shielded slot line for the odd mode.

$$w_2 \simeq a \text{ and } w_2 \simeq 0$$

are compared with those computed for microstrip by [10]. The same type of comparison is also made between the odd mode solution and unilateral finline. Figs. 3 and 4 illustrate the conclusions of the computations for the coupled strip-unilateral finline structure at frequencies $f = 27$ GHz and $f = 40$ GHz and their comparisons for near limiting conditions with microstrip [10], suspended microstrip [10] and unilateral finline [5]. From Fig. 3(a) and (b) showing λ/λ_0, Z_e, α_c, and α_d for the even mode, it is concluded that Z_e, α_d, and λ/λ_0 smoothly approach those of microstrip or suspended microstrip as $w_2 \to 0$ or $w_2 \to a$, respectively. We note that α_c near both limits is significantly larger than that of microstrip, especially as $w_2 \to a$.

In the case of the narrowing slot ($w_2 \to 0$), there will be (for the loss-free case) singularities of field near the two conductor edges that are absent when the slot is absent. In the case of the fins almost vanishing ($w_2 \to a$), this is believed to be a computational artifact, due to the choice of formulation in the plane $y = h$ in terms of electric fields over the slot rather than in terms of conduction current over the fins. Experience [6] has shown this to be the best choice when the slot is smaller than the fin width, which is naturally the region of most practical interest.

Fig. 4(a) and (b) for the odd mode display that there is good agreement between α_c, α_d, Z_o, and λ/λ_o of strip-unilateral finline structure and the values of unilateral finline [5] as the slot becomes larger. The agreement is good for $w_2 > 1.2$ mm, meaning that the strip is not very

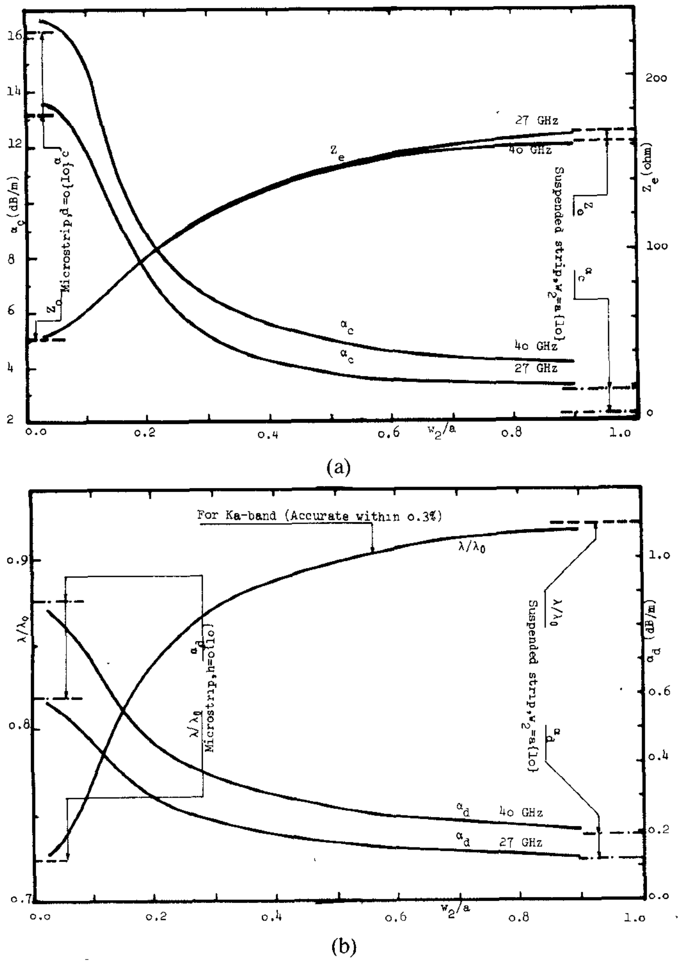

Fig. 3. Even mode parameters of microstrip-unilateral finline built in WG-22 waveguide (Fig. 1(a)) with $a = 1.778$ mm, $h = d = 3.4925$ mm, $\delta = 0.127$ mm, $w_1 = 0.2$ mm, $\epsilon_r = 2.22$ mm, $\tan\delta_t = 2\times10^{-4}$, and $\rho = 3\times10^{-8}$ $\Omega\cdot$m. (a) Characteristic impedance Z_e and conductor loss α_c. (b) Normalized wavelength λ/λ_o and dielectric loss α_d.

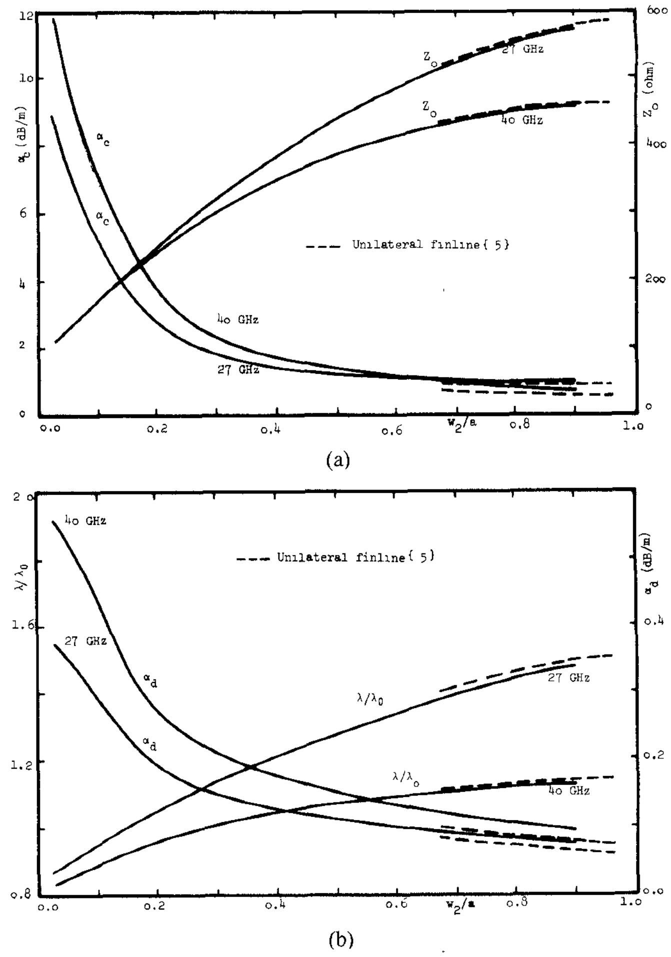

Fig. 4. Odd mode parameters of the microstrip-unilateral finline described in Fig. 3. (a) Characteristic impedance Z_o and conductor loss α_c. (b) Normalized wavelength λ/λ_o and dielectric loss α_d.

effective for this range of w_2. This, however, is understandable because in contrast to the "TEM limit mode," for the odd mode with w_2 chosen very large, the energy does not concentrate about the strip and, therefore, removing the strip would continuously convert the problem to unilateral finline.

With regard to the computation time, it is of the order of 6 s on an IBM 360/65 if $M = N = P = Q = 3$ in (2) to (5). This corresponds to around 0.1 percent accuracy in calculation of the phase constant.

IV. Conclusion

Coupled strip-unilateral finline structures have been analyzed for propagation constant, characteristic impedance, dielectric loss, and conductor loss. The techniques used were the spectral domain hybrid mode analysis to obtain the phase constant and field components for the lossless case, and a perturbation method to determine the dielectric and conductor losses.

Rigorous expansions for the loss-free structure are in terms of the surface currents on the strip together with the electric fields in the finline plane. Concerning the basis functions, they were taken to be trigonometric functions for the bounded field or current, and Legendre polynomials for unbounded field or current. The latter choice is, however, advantageous in calculation of conductor loss. Two examples were treated by the developed method. The first one was used merely for comparison with well-established results of shielded slot and shielded suspended microstrip. The second example, being more realistic, was a coupled strip-unilateral finline built within WG-22 waveguide. In this structure, however, at the appropriate limits very good agreement between the computed parameters and those of microstrip, suspended microstrip, and, unilateral finline was clearly seen. Since the Green's function matrix in the Fourier domain, $[G]$, of the mentioned structure can be used directly for similar cases, its elements have been explicitly given. The manner of the extension of the technique to a general planar structure with multilayer of dielectric and multilayer of conductor was also discussed. Efficiency of the numerical computation is very high and advantage has been taken of the detailed study carried out in [6] for the choice of $[G]$ matrix and of basis function, which strongly influences the computing time.

Appendix

Elements of $[G]$ in (1) are given as follows:

$$G_{m,n} = G'_{m,n}/D, \qquad m,n = 1,2,3,4.$$

$$G'_{1,1} = \beta^2(1 + t_{1,2}^2 t_c)/\cosh(\gamma_{2,n}\delta)$$

and

$$
\begin{aligned}
G'_{1,2} &= \alpha_n \beta^3 k_{1,2}^2 t_c / \cosh(\gamma_{2,n}\delta) \\
G'_{1,3} &= \alpha_n \beta^3 t\left(1 + k_2^2 \gamma_{2,n} t_c\right) \\
G'_{1,4} &= -\beta^2 t\left(k_{2,\beta}^2 + k_2^2 k_{1,\beta}^2 \gamma_{2,n}^2 t_c\right) \\
G'_{2,2} &= \beta^2\left(1 + t_{2,1}^2 t_c\right)/\cosh(\gamma_{2,n}\delta) \\
G'_{2,3} &= -\beta^2 t\left(k_{2,\alpha}^2 + k_2^2 k_{1,\alpha}^2 \gamma_{2,n}^2 t_c\right) \\
G'_{3,1} &= -\beta(E - \alpha_n c' D) \\
G'_{3,2} &= c' k_{3,\beta}^2 D + B \\
G'_{4,1} &= \beta^2 F - c' k_{3,\alpha}^2 D \\
t &= \tanh(\gamma_{2,n}\delta)/\omega\epsilon_2\gamma_{2,n} \\
c &= \coth(\gamma_{1,n}h)/\omega\mu_1\gamma_{1,n} \\
t_c &= t \cdot c \\
c' &= \coth(\gamma_{3,n}d)/\omega\mu_3\gamma_{3,n} \\
E &= -\alpha_n\beta^2\left\{\eta_2 t + c + k_1^2\gamma_{1,n}^2 C_t + \eta_2\left(k_1^2\gamma_{1,n}^2 + k_{1,2}^4\right)T_c\right\} \\
B &= \beta^2\left\{\eta_2 k_{2,\beta}^2 t + k_{1,\beta}^2 c + k_1^2 k_{2,\beta}^2 \gamma_{1,n}^2 C_t \right. \\
&\qquad \left. + \eta_2\left(\alpha_n^2\beta^2 k_{1,2}^2 + k_{2,\beta}^2 t_{2,1}^2\right)T_c\right\} \\
F &= -\left\{\eta_2 k_{2,\alpha}^2 t + k_{1,\alpha}^2 c + k_1^2 k_{2,\alpha}^2 \gamma_{1,n}^2 C_t \right. \\
&\qquad \left. + \eta_2\left(\alpha_n^2\beta^2 k_{1,2}^2 + k_{2,\alpha}^2 t_{1,2}^2\right)T_c\right\} \\
D &= \beta^2\left\{1 + \left(k_2^2\gamma_{1,n}^2 + k_1^2\gamma_{2,n}^2\right)t_c + \left(k_1 k_2 \gamma_{1,n}\gamma_{2,n} t_c\right)^2\right\} \\
k_i^2 &= \omega^2\mu_i\epsilon_i, \qquad \text{for } i = 1,2,3 \\
\eta_2 &= \epsilon_2/\mu_2 \\
T_c &= ct^2 \\
C_t &= c^2 t \\
t_{i,j}^2 &= \alpha_n^2 k_i^2 + \beta k_j^2 - k_i^2 k_j^2 \qquad \text{for } i, j = 1,2 \\
k_{i,\beta}^2 &= k_i^2 - \beta^2, \qquad \text{for } i = 1,2,3 \\
k_{i,\alpha}^2 &= k_i^2 - \alpha^2, \qquad \text{for } i = 1,2,3 \\
k_{1,2}^2 &= k_2^2 - k_1^2 \\
\gamma_{i,n}^2 &= \alpha_n^2 + \beta^2 - k_i^2, \qquad \text{for } i = 1,2,3.
\end{aligned}
$$

In the above expressions, when $\gamma_{i,n}^2 < 0$, all the hyperbolic functions must be replaced by their trigonometric counterparts.

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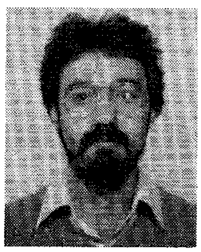

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