SMALL SIZE VCO MODULE FOR 900 MHz BAND USING COUPLED MICROSTRIP-COPLANAR LINES

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Abstract

A hybrid mode analysis of the coupled microstrip-coplanar lines was done, and the transmission and reflection characteristics at each port of the lines were calculated. Applying the coupling characteristics of these lines, a double-sided and miniaturized VCO was made.

Introduction

To miniaturize radio communication equipment to make it portable, one of the most effective ways is to use double-sided integrated circuits. An example of a double-sided IC is the microstrip-slot line directional coupler proposed by F.C.de Ronde in 1970 (1). However, the uses of the coupled microstrip-slot lines are very limited because devices cannot be connected in series to a slot line. We then developed the coupled microstrip-coplanar lines, which can be connected with devices either in series or parallel. This is the desired situation for developing a double-sided IC for use in miniaturized radio equipment. In this paper, we analyze the characteristics of the coupled microstrip-coplanar lines, apply it to a VCO module, and confirm a prototype module through actual operation.

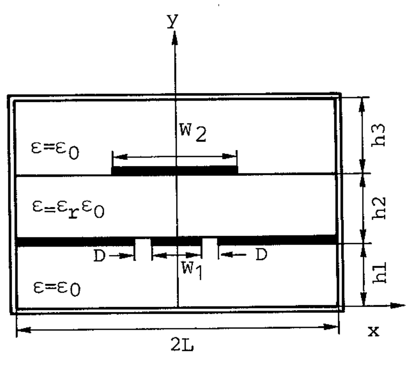

Fig. 1 Coupled microstrip-coplanar lines arranged in a shield case

Eigenmode Analysis of the Coupled Microstrip-Coplanar Lines

Fig.1 shows a shielded three-layer dielectric structure in which coupled microstrip-coplanar lines are arranged. In these coupled lines, the impedance matrix is given as :

$$[\ z\ ] = \begin{bmatrix} z_{11} & z_{12} \\ z_{12} & z_{22} \end{bmatrix}. \qquad (1)$$

The eigenvalues for the impedance matrix [z] and orthogonal eigenvectors corresponding to the eigenvalues are :

$$Z_e = [z_{11}+z_{22}+\sqrt{(z_{11}-z_{22})^2+4z_{12}^2}\ ]\ /\ 2 \qquad (2\text{-a})$$

$$Z_o = [z_{11}+z_{22}-\sqrt{(z_{11}-z_{22})^2+4z_{12}^2}\ ]\ /\ 2 \qquad (2\text{-b})$$

$$X_e = k\begin{bmatrix} 1 \\ p \end{bmatrix},\quad X_o = k\begin{bmatrix} 1 \\ -1/p \end{bmatrix} \qquad (3)$$

$$p = p(z_{11}, z_{22}, z_{12}) = p(Z_e, Z_o, z_{12})$$
$$= [Z_e - Z_o - \sqrt{(Z_e-Z_o)^2-4z_{12}^2}\ ]/2z_{12}. \qquad (4)$$

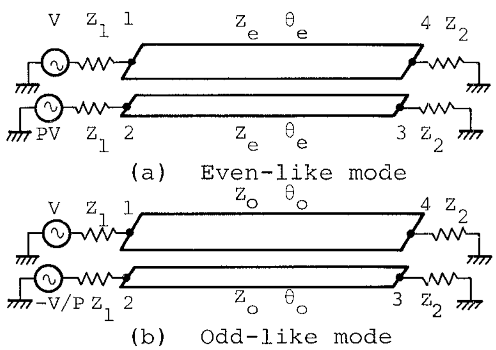

Fig. 2 Eigenmode excitations

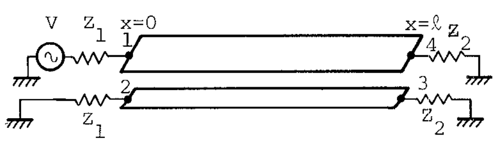

Fig. 3 Excitation of coupled lines

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The eigenvalues are referred to as even-like and odd-like mode impedances Ze and Zo, respectively. The reflection and transmission coefficients under each eigenmode excitation shown in Fig.2 are :

$$R_x = \frac{Z_x(z_2 - z_1) + j(Z_x^2 - x_1 z_2)\tan\theta_x}{Z_x(z_2 + z_1) + j(Z_x^2 + z_1 z_2)\tan\theta_x} \quad (5\text{-a})$$

$$T_x = \frac{2Z_x\sqrt{z_1 z_2}/\cos\theta_x}{Z_x(z_2 + z_1) + j(Z_x^2 + z_1 z_2)\tan\theta_x} \quad (5\text{-b})$$

$$x = e \;,\; o$$

Using the superposition principle, the reflection and transmission coefficients under any type of excitation, such as under the condition shown in Fig.3, are given as :

$$R_1 = 1/(p^2+1)\; R_e + p^2/(p^2+1)\; R_o \quad (6\text{-a})$$

$$R_2 = p/(p^2+1)\; (\, R_e - R_o \,) \quad (6\text{-b})$$

$$T_3 = p/(p^2+1)\; (\, T_e - T_o \,) \quad (6\text{-c})$$

$$T_4 = 1/(p^2+1)\; T_e + p^2/(p^2+1)\; T_o \;. \quad (6\text{-d})$$

Calculation of Parameters

To obtain dispersion characteristics of the type of structure shown in Fig.1, we employ the "spectral domain" method (2),(3),(4),(5). Let the electrical and the magnetic potential functions be $\phi_i^e(x,y)$ and $\phi_i^h(x,y)$ (i=1,2,3), respectively. These scalar functions satisfy the Helmholtz equation. To transform the Helmholtz equation into the spectral domain, we introduce the following Fourier integral :

$$\phi_i^{e,h}(k_n,y) = \int_{-L}^{L} \phi_i^{e,h}(x,y)\,\mathrm{EXP}(jk_n x)\,dx. \quad (7)$$

Solving the transformed Helmholtz equation, the following potential functions are obtained :

$$\phi_1^e(k_n,y) = A\sinh\gamma_1 y F_e(k_n)$$

$$\phi_2^e(k_n,y) = [B\sinh\gamma_2(h_1+h_2-y) + C\cosh\gamma_2(h_1+h_2-y)]F_e(k_n)^2$$

$$\phi_3^e(k_n,y) = D\sinh\gamma_3(h_1+h_2+h_3-y)F_e(k_n)$$

$$\phi_1^h(k_n,y) = E\cosh\gamma_1 y F_o(k_n) \quad (8)$$

$$\phi_2^h(k_n,y) = [F\cosh\gamma_2(h_1+h_2-y) + G\sinh\gamma_2(h_1+h_2-y)]F_o(k_n)^2$$

$$\phi_3^h(k_n,y) = H\cosh\gamma_3(h_1+h_2+h_3-y)F_o(k_n)$$

where $F_e(k_n) = \int_{-L}^{L}\cos(k_n x)\,\mathrm{EXP}(jk_n x)\,dx$

$$F_o(k_n) = \int_{-L}^{L}\sin(k_n x)\,\mathrm{EXP}(jk_n x)\,dx$$

$$k_n = (n+0.5)\pi/L$$

Taking into account the boundary conditions at the interfaces $y=h_1$ and $y=h_1+h_2$, the unknown constants are determined and the following set of coupled equations can be obtained :

$$\begin{bmatrix} G11 & G12 & G13 & G14 \\ G21 & G22 & G23 & G24 \\ G31 & G32 & G33 & G34 \\ G41 & G42 & G43 & G44 \end{bmatrix} \cdot \begin{bmatrix} I_{x1} \\ I_{z1} \\ I_{x2} \\ I_{z2} \end{bmatrix} = \begin{bmatrix} E_{z1} \\ E_{x1} \\ E_{z2} \\ E_{x2} \end{bmatrix} \quad (9)\;.$$

In Eq.(9), Ix1 and Iz1 are the transformed current densities and Ex1 and Ez1 are the electric fields at $y=h_1$, while Ix2, Iz2, Ex2 and Ez2 are the same parameters at $y=h_1+h_2$. To solve Eq.(9), we expand Ixi, Izi (i=1,2) into a finite series using a complete set of basis functions, as shown in the following equations :

$$I_{xi} = \sum_l \begin{pmatrix} a_l \\ b_l \end{pmatrix} \sin(l\pi x/W_i) \;:\text{even mode} \quad (10\text{-a})$$

$$I_{zi} = \sum_l \begin{pmatrix} c_l \\ d_l \end{pmatrix} P_{2(l-1)}(x/W_i) \quad (10\text{-b})$$

and

$$I_{xi} = \sum_l \begin{pmatrix} e_l \\ -f_l \end{pmatrix} \sin(l\pi x/W_i) \;:\text{odd mode} \quad (11\text{-c})$$

$$I_{zi} = \sum_l \begin{pmatrix} g_l \\ -h_l \end{pmatrix} P_{2(l-1)}(x/W_i) \quad (11\text{-d})$$

Where Pn(x/Wi) (i=1,2) is Legendre's polynomial. Applying Galerkin's method and Parseval's identity, the following equation can be obtained :

$$\begin{vmatrix}
\sum_1^K K^{1,1}_{1,1} & \sum_1^L K^{1,2}_{1,1} & \sum_1^M K^{1,3}_{1,1} & \sum_1^N K^{1,4}_{1,1} \\
\sum_1^K K^{1,1}_{1,K} & \sum_1^L K^{1,2}_{1,K} & \sum_1^M K^{1,3}_{1,K} & \sum_1^N K^{1,4}_{1,K} \\
\sum_1^K K^{2,1}_{1,1} & \sum_1^L K^{2,2}_{1,1} & \sum_1^M K^{2,3}_{1,1} & \sum_1^N K^{2,4}_{1,1} \\
\sum_1^K K^{2,1}_{1,L} & \sum_1^L K^{2,2}_{1,L} & \sum_1^M K^{2,3}_{1,L} & \sum_1^N K^{2,4}_{1,L} \\
\sum_1^K K^{3,1}_{1,1} & \sum_1^L K^{3,2}_{1,1} & \sum_1^M K^{3,3}_{1,1} & \sum_1^N K^{3,4}_{1,1} \\
\sum_1^K K^{3,1}_{1,M} & \sum_1^L K^{3,2}_{1,M} & \sum_1^M K^{3,3}_{1,M} & \sum_1^N K^{3,4}_{1,M} \\
\sum_1^K K^{4,1}_{1,1} & \sum_1^L K^{4,2}_{1,1} & \sum_1^M K^{4,3}_{1,1} & \sum_1^N K^{4,4}_{1,1} \\
\sum_1^K K^{4,1}_{1,N} & \sum_1^L K^{4,2}_{1,N} & \sum_1^M K^{4,3}_{1,N} & \sum_1^N K^{4,4}_{1,N}
\end{vmatrix} = 0 \quad (12)$$

where $K^{i,j}_{l,m} = \sum_{n=0}^{\infty} I_{\substack{x\\z},l}\, Gij\, I_{\substack{x\\z},m}$.

From Eq.(12), the propagation constants can be calculated.

Fig.4 shows electric fields under both eigenmode excitations. From the configuration of the electric fields, we define the characteristic impedances corresponding to the eigenmode excitations, such as :

$$Z_e = 4P_{ave} \;/\; I(1+p)^2 \qquad (13\text{-a})$$

$$Z_o = V(1+1/p)^2 \;/\; 4P_{ave} \qquad (13\text{-b})$$

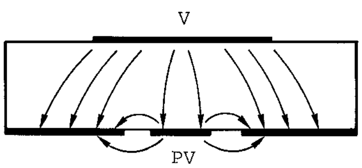
$$P_{ave} = 1/2\mathrm{Re}\iint E \times H^{*}\cdot a \; dxdy \qquad (13\text{-c})$$

$$I(1+p) = \int_{-W2}^{W2} I_{z2}(x)\; dx \qquad (13\text{-d})$$

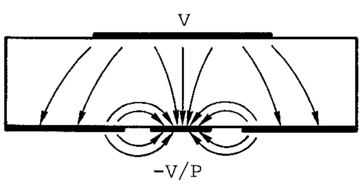
$$V(1+1/p) = \int_{h1}^{h1+h2} E_y(0,y)\; dy \; \cdot \qquad (13\text{-e})$$

As hybrid mode calculation for z12 is difficult, we determine it using TEM approximation.

Fig. 5 shows an example of the coupling characteristics calculated using this method.



(a) Even-like mode



(b) Odd-like mode

Fig. 4 Electric field distribution

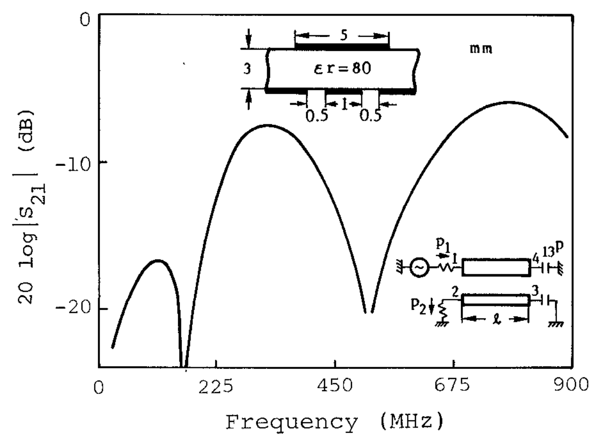


Fig. 5 Calculated coupling characteristics

VCO Using Coupled Microstrip-coplanar Lines

We fabricated a small size VCO for a 900 MHz band using the coupled microstrip-coplanar lines. The signal generated by the oscillator on the main plane of the substrate is fed to the buffer amplifier on the opposite side. This is done by using the coupling characteristics between the $\lambda g/4$ microstrip resonator and the coplanar line which face each other through the substrate. A hybrid VCO circuit made for trial is shown in Fig.6. With dimensions of 15x22x12 mm, its output power was 0 dBm at Vcc=5.5 V. The measured characteristics of oscillating frequency versus tuning voltage were about 7 MHz/V, while $C/N \geq 70$ dBc (typically 73 dBc at offset frequency=25 kHz, bandwidth= 15 kHz). Fig.7 shows the oscillating characteristics of the VCO fabricated for trial.

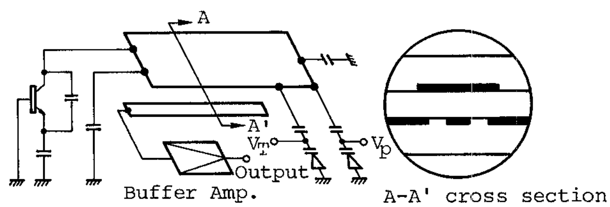


Fig. 6 Trial VCO circuit using coupled microstrip-coplanar lines

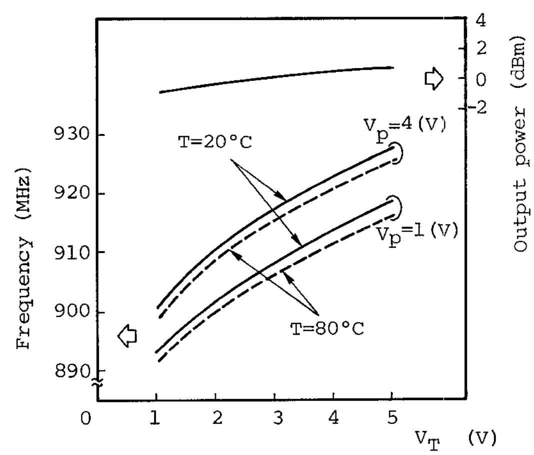


Fig. 7 Oscillating characteristics of VCO fabricated for trial

Conclusions

A new coupling configuration incorporating two transmission lines to form both sides of a substrate, termed coupled microstrip-coplanar lines, was developed. A hybrid mode analysis of the coupled microstrip-coplanar lines was done, and the coupling characteristics were applied to a VCO module to facilitate miniaturization of components. Judging by the VCO module's performance and size, we concluded that the method for miniaturization mentioned in this study is effective.

Acknowledgement

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References

(1) F.C.de Ronde, "A New Class of Microstrip Directional Couplers," in Proc. IEEE Int. Microwave Symp., pp.184-186, 1970.

(2) T.Itoh and R.Mittra, "Spectral-Domain Approach for Calculating the Dispersion Characteristics of Microstrip Lines," IEEE Trans. M.T.T., vol.MTT-21, pp.496-499, July 1973.

(3) T.Itoh and R.Mittra, "A Technique for Computing Dispersion Characteristics of Shielded Microstrip Lines," IEEE Trans. M.T.T., vol.MTT-22, pp.896-898, Oct. 1974.

(4) J.B.Davis and D.Mirshekar-Syahkal, "Spectral Domain Solution of Arbitrary Coplanar Transmission Line with Multilayer Substrate," IEEE Trans. M.T.T., vol.MTT-25, pp.143-146, Feb. 1977.

(5) D.Mirshekar-Syahkal and J.B.Davis, "Accurate Solution of Microstrip and Coplanar Structures for Dispersion and for Dielectric and Conductor Losses," IEEE Trans. M.T.T., vol.MTT-27, pp.694-499, July 1979.