Combination Dielectric Resonator, Power Combiner, and Antenna

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***Abstract*—A rectangular dielectric resonator housed in a cutoff parallel-plate waveguide is used both as a radiating element and microwave power combiner. The resonator is excited by using tuned electrically short monopole antennas to induce a longitudinal electric operating mode. The resonator is then used in conjunction with free-running oscillators in order to provide, via mutual injection locking, stable in-phase power combining. Furthermore, the resonator is arranged such that one of its faces radiates a portion of the power-combined signal. Since the resonator is housed in a cutoff waveguide, the cross-polarization radiation from the antenna is suppressed. It was found that, for a single element, a gain in the azimuthal plane of 5 dB could be achieved and, for a two-element array, a gain of 7 dB was obtained with better than −25-dB cross polarization for each case. The oscillator power-combining efficiency for a single-element antenna (two oscillators) was 91%, and the spatial power-combining efficiency for a two-element antenna array, (four oscillators) was found to be 90%. In addition, it is shown that the presence of the dielectric inserts in conjunction with coupled oscillator dynamics provides moderate overall oscillator phase noise improvement.**

***Index Terms*—Dielectric resonator, spatial power combiner.**

I. Introduction

DIELECTRIC waveguides, especially nonradiative dielectric (NRD) waveguide, have advantages in the application of millimeter-wave integrated circuits [1]–[3]. Basically, the NRD waveguide, shown in Fig. 1, resembles a modification of the structure of Tischer's H-guide [4].

The typical NRD waveguide, shown in Fig. 1, supports hybrid modes that have longitudinal electric- or magnetic-field component along the propagation direction z, where the waveguide is assumed for the present to be infinitely long. These modes can be categorized into longitudinal electric modes (LSE modes) and longitudinal section magnetic modes (LSM modes). It is known that the LSE_{10} mode is the lowest frequency zero cutoff mode, the LSE_{11}, LSM_{10}, LSM_{11} modes are the higher order modes, and they all possess the characteristics of suppressing transverse radiation [1]–[3]. The lowest order mode LSM_{10} has maximum electric field along the center of the dielectric material and can be excited by inserting a short monopole placed along the y-direction into the dielectric material of Fig. 1. In this paper, it is shown that, with careful choice, the positions of the monopoles inserted into the dielectric block (which, in our paper, is made to have finite length in the z-direction) and their subsequent connection to free-running oscillators enables efficient microwave power combination to be realized. In addition, mutual coupling between oscillators by virtue of the principal mode excited by the choice of configuration used ensures that the oscillators become frequency-locked together. Furthermore, the presence of multiple coupled oscillators and the dielectric insert forming the resonator block act to improve oscillator output stability without the need for individual separate dielectric-resonator oscillator (DRO) stabilization of each oscillator. In our paper, the load for the power combiner thus formed is free space since the dielectric face positioned at $z = 0$ in conjunction with the upper and lower metal plates, as shown in Fig. 2, also serves the additional purpose of operating as an antenna element.

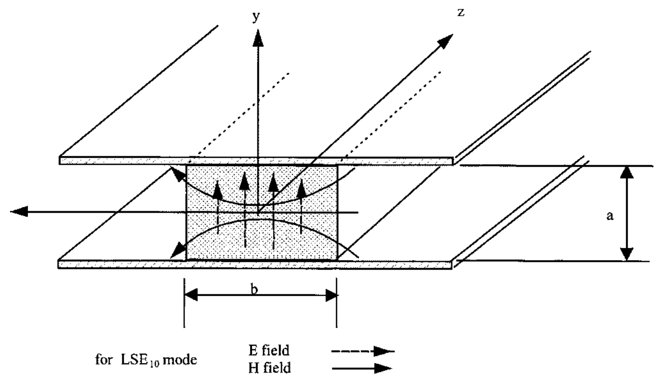

Fig. 1. LSE_{10} mode in NRD waveguide.

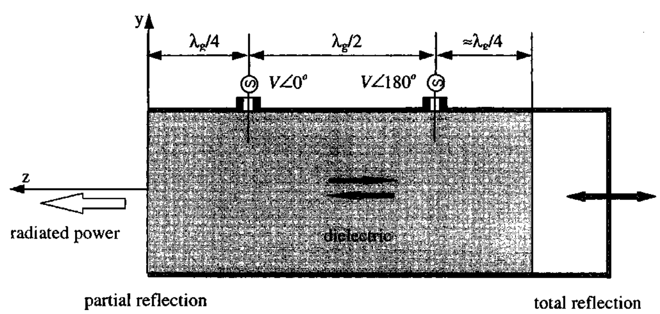

Fig. 2. Resonator elevation.

II. Operation of Dielectric Waveguide Combiner/Antenna Element

It is known that by solving a set of eigenvalue equations established through matching the electromagnetic fields at horizontal

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TABLE I
GUIDE WAVELENGTH VERSUS DIELECTRIC WIDTH b AND HEIGHT a

Given NRD width b/λ_o	calculated λ_{g1}/λ_o for $\varepsilon_r = 2.56$	calculated λ_{go}/λ_o for $\varepsilon_r = 2.56$	single-mode operating NRD height a/λ_o
0.30	0.99	0.84	0.42
0.37	0.98	0.74	0.37
0.40	0.96	0.73	0.36

and vertical dielectric boundaries with air, the guide wavelength in the dielectric is determined [1]–[3] as

$$\lambda_{gn} = \frac{\lambda_o}{\sqrt{\varepsilon_r - (q_n \lambda_o / 2\pi)^2}}. \tag{1}$$

Here, λ_{gn} is the guide wavelength in the dielectric, λ_o is the free-space wavelength, ε_r the relative permittivity of dielectric, and q_n is the nth solution of the following characteristic equations:

$$\frac{q_n}{\varepsilon_r} \tan\left(\frac{q_n b}{2}\right) - \sqrt{(\varepsilon_r - 1)\left(\frac{2\pi}{\lambda_o}\right)^2 - q_n^2} = 0, \qquad \text{for even } n \tag{2a}$$

$$\frac{q_n}{\varepsilon_r} \cot\left(\frac{q_n b}{2}\right) + \sqrt{(\varepsilon_r - 1)\left(\frac{2\pi}{\lambda_o}\right)^2 - q_n^2} = 0, \qquad \text{for odd } n. \tag{2b}$$

For given dielectric constant ε_r, free-space wavelength λ_o and size b, (2a) and (2b) allow λ_{gn} to be determined from (1). Single-mode operation in the y-direction is determined by height a so that the higher order modes are cutoff; the condition for this to occur [3] is expressed by (3) as

$$\lambda_o, \lambda_{g1} > 2a > \lambda_{go} \tag{3}$$

where λ_{go} and λ_{g1} are the guide wavelengths of the fundamental and second-order modes, respectively. The free-space wavelength is included in (3) to ensure suppression of radiated waves in the x-direction within the parallel-plate waveguide structure. It is this property that gives the arrangement low cross polarization when the exposed dielectric face in Fig. 2 is allowed to radiate into free space.

For an NRD waveguide using polystyrene as the dielectric insert $\varepsilon_r = 2.56$ at $\lambda_o = 30$ cm (RF frequency at 1 GHz), the guide wavelengths of the fundamental and second-order modes solved using (2a) and (2b) are listed in Table I for different dielectric widths b. The required minimum dielectric height a is obtained from (3); hence, we select $a = b = 0.37\ \lambda_o = 11.1$ cm.

As discussed below for in-phase power combining, the length of the resonator should be chosen to be an integral number of half guide wavelengths in order to satisfy the wave propagation boundaries for resonance along the z-direction. In our paper, we selected $z = \lambda_{g0} = 22.2$ cm and neglected end fringing effects. The top and bottom metal plates were allowed to extend 30 cm in the positive and negative x-directions, respectively (Fig. 1).

III. Single-Element Considerations

By inputting signals from two suitably tuned short monopoles mounted through the upper metal plate of the resonator, a partial standing wave was excited and power combination was realized. By terminating the dielectric slab in the z-direction (Fig. 2) it is possible to construct a resonator.

The power carried by the dipoles is absorbed with the correct phase relationship for reinforcement by the induced wave as it moves both in the positive and negative z-directions. The backward wave is reflected into the dielectric by the earthed metal plane, and the radiating power is produced from the forward wave (Fig. 2). Any mismatch between the forward wave and free space that happens at the output plane of the dielectric antenna can be minimized by connecting a transformer waveguide to the resonator [7] or by using a matching flange along the open face of the waveguide.

The air gap between the back earth plane and one end of the dielectric block is made variable using a short circuit. Its distance from the near face of the dielectric block is initially selected to be equal to half of the dielectric height a in order to cutoff radiation effectively along the backward direction and to minimize the earth plane influences on the LSE mode operation. In Fig. 2, since the resonator element is approximately one guide wavelength long, monopole two when placed in the upper metal plate must have a 180° phase delay introduced in series with its connection to oscillator 2 relative to the connection of monopole 1 with respect to oscillator 1. For now, we have assumed that both oscillators are operating in-phase. This arrangement is necessary in order to phase the excitation sources such that power combining occurs in the leaky cavity formed by the structure in Fig. 2 [5].

The relative E_y electric-field amplitude distribution in the x–y-plane at $z = 3$ mm is shown in Fig. 3 and is compared with results predicted from HFSS [6] together with measured data obtained using a short balanced dipole placed 3 mm from the front face of the open dielectric place and parallel to the y-axis. Both sets of results have been normalized to unity at the center of the dielectric face $x = 0$ (Fig. 1). To perform this experiment, a single source was equally split and fed directly to probe 1 and

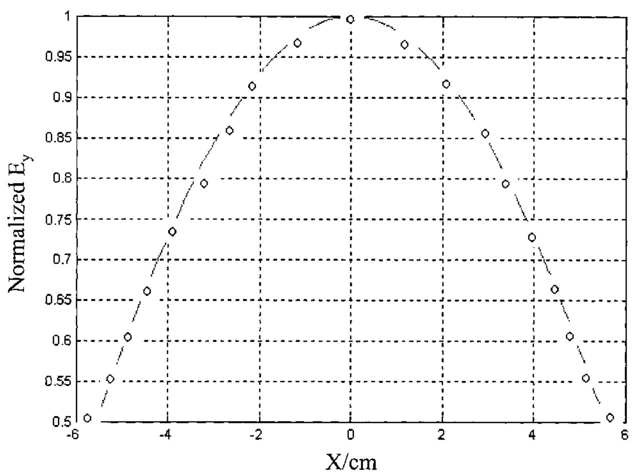

Fig. 3. Simulated (-) and measured (0) E_y field distribution on the output face of the resonator block.

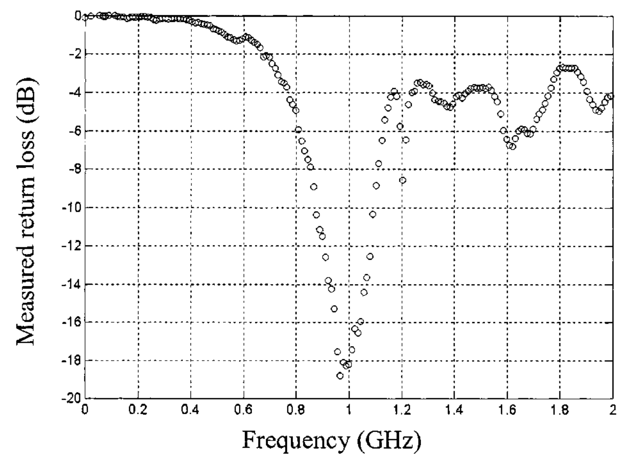

Fig. 4. *In situ* measured monopole return loss.

an additional 180° phase shift was introduced into the feed to probe 2.

As shown above, both the predicted results from HFSS and the measured data obtained by a short dipole antenna probe indicate that the electric field reaches its highest value at the center of the dielectric block, the electric field at the edge of dielectric block $x = \pm 5.55$ cm is 3 dB less than that at center position. In the free-space region inside the waveguide, measured at $z = -3$ mm on either side of the resonator at $|x| = 5.8$ cm, the measured electric field was reduced by more than -10 dB when compared with the field simulated in the dielectric block. In addition, the measured E_x component was 25 dB less than the E_y component, which supports the cutoff waveguide concept suggested above.

To input microwave power into the resonator antenna, the input impedance of the short monopoles used in Fig. 2 must be matched to the leaky-cavity resonator impedance. The electrically short monopole antennas used to couple power to this mode were designed according to the procedure given in [8]. In our experiments, the diameter of monopole was fixed at 0.006 λ_g and the length of monopole was 0.18 λ_g. The return loss for the monopole when inserted into the dielectric was measured. Fig. 4 shows the measured return loss versus frequency. Due to the presence of the back plate in Fig. 2, (1) becomes approximate when applied to the calculation of the z-dimension of the resonator. Therefore, due to extra fringing fields at the rear of the resonator and also at its front face, the actual structure was resonant at 998 MHz, as shown in Fig. 4. As seen here, the typical value of return loss at 998 MHz was -19 dB, the estimated loaded Q of the structure when operating into free space was ten.

In addition, the measured mutual coupling between adjacent monopoles placed in a single dielectric block at resonance is -10 dB, and is capacitive as required for single-frequency entrainment of mutually injection locked oscillators [9].

As a result of the above discussions on power-combining requirements, we come to conclusion that: 1) the relative phase of the excitation signals on adjacent monopoles must be 180° in order to form constructive power combining in the dielectric region and 2) suitably phased excitation between adjacent dipoles should act to perform destructive combining.

To test the principle for a single resonator, the following experiment was performed. For the different two-monopole excitation cases, the measured relative electric-field intensity (normalized to case (ii) in Table II) at the center of output window plane ($x = 0$, $y = 0$, $z = 3$ mm) are summarized in Table II.

Here, the quoted E_y component was obtained by using a single source fed to each monopole via a power splitter. This data shows that destructive power combining is occurring at -27 dB relative to the constructive configuration. The power-combining efficiency obtained is about 91%. When the single oscillator used in Table II was replaced by two free-running oscillators, each producing the same output power, the power combining results in Table III were obtained using the same normalization method shown in Table II.

This time when the oscillators acquire frequency entrainment by virtue of mutual locking [10]–[12], they do not run exactly in phase due to the static phase error that exists between them. Consequently, a reduced power-combining efficiency of 72% is obtained in this scenario.

During the experiment, a typical locked signal bandwidth of 0.3 MHz was observed and harmonic components were seen to be better than -20 dB relative to the locked center-frequency component.

IV. Dual Element

By arranging two resonators side by side, an array can be formed, as shown in Fig. 5. Here, the coupling between monopoles 1 and 4 is given in Table IV as function of element separation; also given is the locking bandwidth B of the system. As can be seen at $d/\lambda_g > 0.1$, the coupling is less than -10 dB and the locking bandwidth is less than 0.14 MHz. The levels of coupling in Table IV are comparable with those reported in [3] for a double-strip H-guide structure.

It was found that when the array elements are placed very close together ($d/\lambda_g < 0.02$), the resonant bandwidth was reduced and excitation of higher order modes becomes a distinct

TABLE II

Input conditions at dipole 1 and dipole 2 (input signal amplitude(relative phase))	Relative E_y
(i) 0.0 dBm (0° at dipole 1); 0.0 dBm (0° at dipole 2)	-27.0 dB (destructive)
(ii) 0.0 dBm (0° at dipole 1); 0.0 dBm (180° at dipole 2)	0.0 dB (constructive)
(iii) -50 dBm dipole 1;0.0 dBm at dipole 2	-5.6 dB
(iv) 0.0 dBm at dipole 1; -50 dBm dipole 2	-5.6 dB

TABLE III

Input conditions at dipole 1 and dipole 2 (input signal amplitude (relative phase))	Relative E_y
0.0 dBm (0° at dipole 1); 0.0 dBm (0° at dipole 2)	-27.0 dB
0.0 dBm (0° at dipole 1); 0.0 dBm (180° at dipole 2)	0.0 dB
-50 dB dipole 1;0.0 dBm at dipole 2	-4.6 dB
0.0 dBm at dipole 1; -50 dBm dipole 2	-4.8 dB

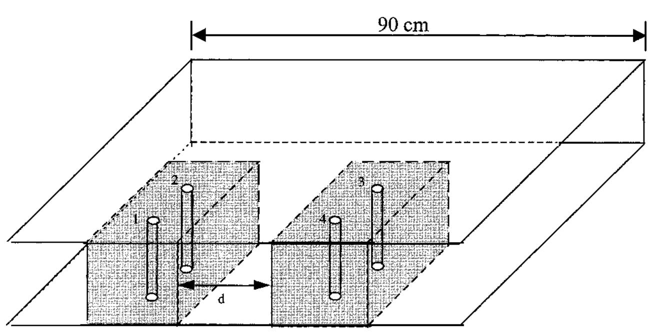

Fig. 5. Two-element array.

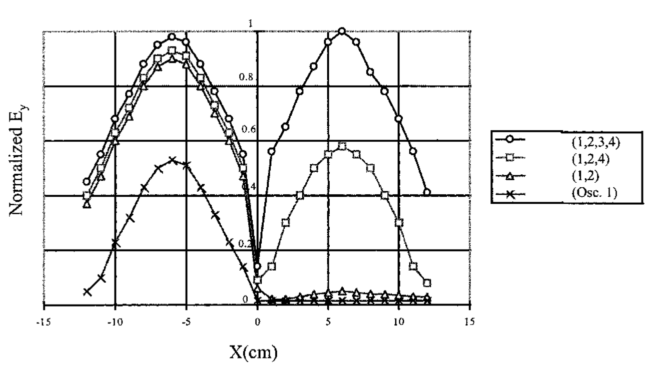

Fig. 6. Measured near-field electric field for two elements.

possibility; a similar finding was reported in [13] for a cylindrical dielectric resonator antenna.

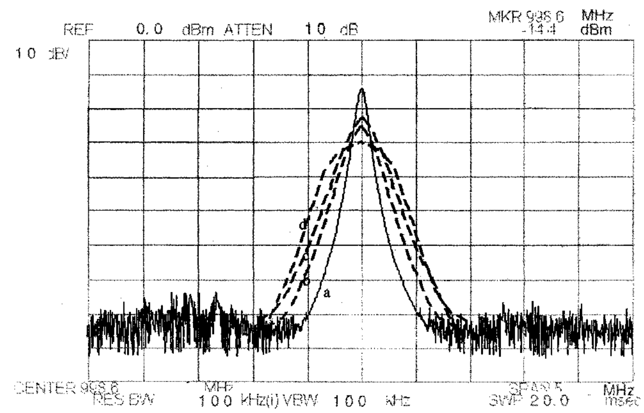

Fig. 7. Typical output spectrum for different oscillator configurations (RB = 100 kHz, VB = 100 kHz, center frequency = 998.6 MHz, frequency span = 5 MHz). Curve (a) is for four-oscillator operation. Curve (b) is for three-oscillator operation. Curve (c) is for two-oscillator operation. Curve (d) is for one-oscillator operation. The vertical scale is 10 dB/div.

The frequency locking range for four-monopole excitation was found to be 0.16 MHz at an array element separation distance of $d = 0.045\ \lambda_g$, at which position the relative phase differences between monopole pairs (1, 2) and (3, 4) were measured to be 180°, while between monopole pairs (1, 4) and (2, 3), they were zero.

The typical measured normalized relative electric-field distribution along the x-direction on the $z = 3.3$ mm plane is shown in Fig. 6 for different excitation configurations. Again, these results were obtained using a single source and power splitter, as described previously for the single-element case results presented in Fig. 3.

Here it can be seen that, as excitation is applied to all four monopoles, power combining is occurring and in-phase signals are available for power combining in the space in front of the antenna array, i.e., constructive spatial power combining can now occur.

The associated normalized frequency spectrums for each different oscillator configuration tested are shown in Fig. 7. As can be seen from Fig. 7, the spectral purity of the output from the resonator array is enhanced, owing to the power-combining process leading to mutual locking over progressively narrower locking bandwidths. Here, when the monopoles are excited by

TABLE IV

d/λ_g	0.01	0.02	0.04	0.06	0.08	0.10	0.12	0.14	0.16
$\|S_{14}\|$	0.201	0.169	0.158	0.138	0.115	0.096	0.068	0.045	0.023
B (MHz)	0.10	0.14	0.16	0.15	0.14	0.14	0.14	0.11	0.09

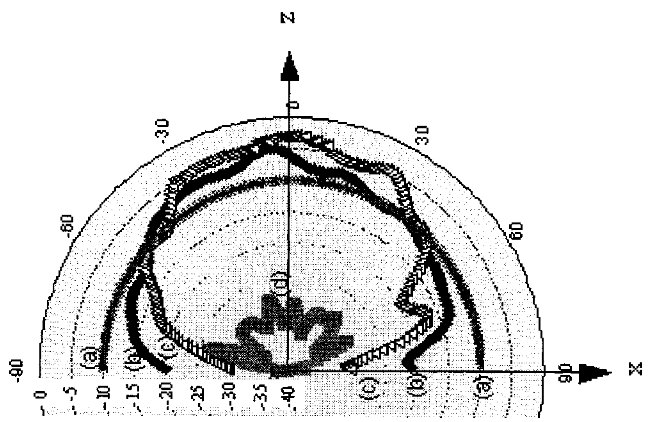

Fig. 8. Measured far-field radiation patterns on the XZ-plane. (a) E_y for a half-wavelength dipole antenna. (b) E_y for a single resonator. (c) E_y for a 1×2 resonator array with element separation 0.4 λ_0. (d) E_x pattern for a single element.

oscillators, these will mutually couple to each other and injection lock. In doing so, the presence of multiple coupled oscillators [14] and the dielectric resonators act to provide oscillator stabilization by virtue of the reciprocal coupling network created. Consequently, a reduction in phase noise occurs. Further research on mutual coupling mode mechanisms and resultant unloaded Q-factor determination is needed to completely understand the frequency-locking mechanism and phase-noise improvement obtained for the configuration presented here.

V. Far-Field Radiation Pattern

Fig. 8 shows measured far-field radiation E-field patterns for the single- and two-element arrangements discussed in this paper. For a VSWR of less than 1.5 (-14-dB return loss), a single element has a bandwidth of 15%; its gain referenced to a standard $\lambda_o/4$ length dipole antenna [see Fig. 8(a)] was found to be 5.0 dB and a half-power beam width of 60° was obtained on the principle E-plane [see Fig. 8(b)]. A half power beam width of 40° was obtained on the principle E-plane after spacial combining of the frequency locked 1×2 array with element separation distance of 0.4 λ_o [see Fig. 8(c)]. The gain of antenna array referenced to a $\lambda/4$ length dipole antenna was found to be 7.0 dB and cross-polarization levels were below -25 dB [see Fig. 8(d)] for both single- and dual-element arrangements. The spatial power coupling efficiency of the two-element arrangement was estimated to be 90%.

VI. Conclusions

A dielectric resonator serving the multiple purposes of microwave power combining, oscillator mutual coupling, oscillator phase noise reduction, and antenna element has been designed, fabricated, and characterized. Spatial power-combining efficiency of 90% at a frequency of 998 MHz was achieved in a two-element resonator array. The resulting antenna E-field radiation pattern exhibits approximately for a single-element 5-dB gain and for a two-element 7-dB gain. In all cases, the measured cross-polarization levels were less than 25 dB. The structure demonstrated here could be miniaturized for use at millimeter-wave frequencies as a source in transmitter applications where enhanced output power is required.

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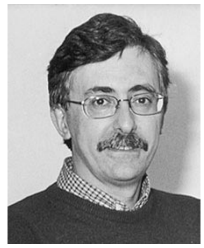

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