A Periodic Second Harmonic Spatial Power Combining Oscillator

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***Abstract*—In this paper we present a periodic second harmonic spatial power combining oscillator. The power combining is achieved by phase locking the oscillators at the fundamental frequency and combining the second harmonic power in space through an array of microstrip patch antennas. The effect of moding and multiple device-circuit interaction is investigated. This circuit is planar, and therefore simplifies the design of monolithic circuits. In this work, *X*-band Gunn diodes are used for the purpose of demonstration.**

Introduction

At MILLIMETER and sub-millimeter wave frequencies solid state devices have limited capability to produce microwave energy; the efficiency of solid state devices drops dramatically. In order to obtain higher power, it is very desirable to combine the power generated from many single solid state oscillators [1]. In addition, there is always a limit to the frequency at which a solid state oscillator can operate. In that limit, the negative resistance of the device becomes so small that it can not compensate for the resonator's loss. In this case power combining alone cannot help. One way to generate millimeter, sub-millimeter wave and terahertz frequencies is to use two terminal solid-state devices as harmonic generators. Even though solid state devices do not exhibit negative resistance above a certain range of frequencies, higher harmonics are generated because of device nonlinearities. Through proper circuit design harmonic generation can be enhanced and then filtered out for use. One problem with this method is the low conversion efficiency. Here again we can take advantage of different power combining techniques [2]. In general at millimeter wave frequencies and above, spatial power combining techniques are preferred since guiding structures are lossy [3], [5].

In this paper we present a periodic second harmonic spatial power combining structure. A periodic structure designed for fundamental power combining is described in [4]. The circuit we discuss here dissipates very low power at the fundamental frequency of operation, making second harmonic generation more efficient. In this circuit devices are connected periodically to a microstrip line structure. Phase locking is accomplished through the strong interaction between negative resistance devices at the fundamental frequency. In this respect the structure is completely planar and there is no need for an external resonator such as a Fabry–Perot resonator to ensure phase locking [5], [6], thus this circuit is suitable for use in monolithic integrated circuits. An antenna array can be designed to radiate only the desirable harmonic frequency. Also this circuit can be modified for use as a second harmonic quasi-optical transceiver, with an array of microstrip patch antennas used for transmitting and receiving purposes. For receiving, the RF energy captured by the antenna array is mixed down with the second harmonic generated from the active devices and the IF signal output is through a low-pass filter.

First, we will describe the multi-device circuit's modes of operation. Then, we will investigate the second harmonic generation using a negative resistance device such as a Gunn diode and how to maximize the power generated at the second harmonic by a single device. Finally, we will consider the design procedure for the complete circuit.

Circuit Description and Theory

The periodic second harmonic spatial power combining oscillator is shown in Fig. 1. A microstrip transmission line is loaded periodically with four negative resistance devices with the distance between them equal to approximately a half guide wavelength at the fundamental frequency of oscillation (f_1). The guide wavelength at the fundamental frequency is λ_{g1} and at the second harmonic is λ_{g2}.

For multiple device oscillators, the potential number of modes of operation is at least equal to the number of devices used in the circuit, due to the fact that a single device oscillator can have more than one mode of operation. Here we will show that in our circuit only one mode of oscillation can exist at the design frequency. This is true at both the fundamental frequency of oscillation as well as the second harmonic. We use a method similar to that of Kurokawa [7] and Peterson [8] to address moding effects in harmonic power combining oscillators. The major

Manuscript received May 21, 1991; revised September 9, 1991. This work was supported by the Joint Services Electronics Program under Contract AFOSR 49620-89-C-0044, by the Army Research Office under contract DAAL03-88-K-0005, and by the Texas Advanced Technology Program.

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IEEE Log Number 9106782.

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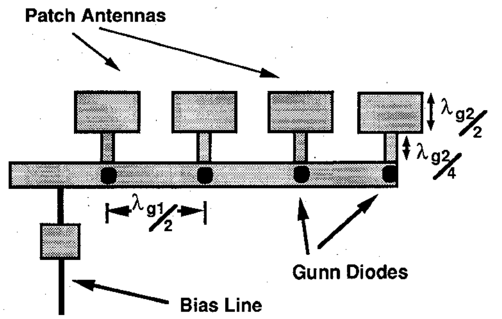

Fig. 1. Diagram of a four diode spatial second harmonic power combiner. λ_{g1} and λ_{g2} are guide wavelengths at fundamental and second harmonic, respectively.

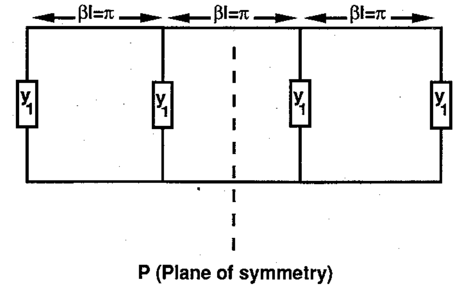

Fig. 2. Equivalent circuit for the four diode harmonic combiner at the fundamental frequency. The admittance of the microstrip patches and feed lines is represented by y_1.

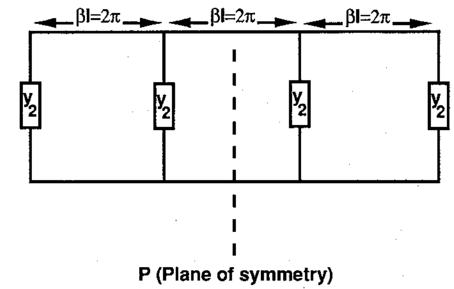

Fig. 3. Equivalent circuit for the four diode harmonic combiner at the second harmonic. The admittance of the microstrip patches and feed lines is represented by y_2.

difference between our circuit and the one analyzed in references [7] and [8], is that our circuit possesses certain symmetries only at frequencies at which the distance between active devices is an integer multiple of a half guide wavelength.

Let $\boldsymbol{Z}$ be the impedance matrix of the circuit as seen from the device ports, and $\overline{\mathbf{Z}}$ be a diagonal matrix with each element equal to the negative of the device impedance. Following Kurokawa, it is found that a set of equations

$$X_n^T \cdot (\lambda_n I - \overline{\mathbf{Z}})i = 0 \qquad n = 1, \cdots, N \tag{1}$$

must be satisfied in order to fulfill the oscillation condition. X_n and λ_n are the n$^{\text{th}}$ eigenvector and eigenvalue of the matrix $\boldsymbol{Z}$, N is the number of devices, and i is a vector containing the currents at each device port.

We will now restrict our attention to a four diode example. The equivalent circuit for the four diode combiner is shown in Figs. 2 and 3. The admittances of the microstrip patches with the feed lines (lines connecting the patch antennas to the main structure) at the fundamental and second harmonic frequencies are y_1 and y_2, respectively. Let $\boldsymbol{Z}$ be the impedance matrix of the oscillator circuit looking in from the device ports. For the four diode power combiner, the plane of symmetry "P" is located between the second and the third devices (Figs. 2 and 3). Due to this symmetry, and to the reciprocity relation $z_{mn} = z_{nm}$, it can be seen that the impedance matrix of the combiner circuit has the following form

$$\boldsymbol{Z} = \begin{bmatrix} z_{11} & z_{12} & z_{13} & z_{14} \\ z_{12} & z_{22} & z_{23} & z_{24} \\ z_{13} & z_{23} & z_{22} & z_{12} \\ z_{14} & z_{13} & z_{12} & z_{11} \end{bmatrix}. \tag{2}$$

The impedance matrix has the same form both at the fundamental and at the second harmonic. The eigenvectors and eigenvalues of the above matrix can now be determined. In [7], it was concluded that if the devices all present the same impedance, any possible solution for i must be an eigenvector and the device impedance must be the negative of the corresponding eigenvalue. However, in general in the large signal case, the device impedances are a function of amplitude and thus may differ between the various devices, even if the physical devices are identical, so a slightly different approach must be taken.

First we consider only the intended fundamental frequency of operation, to insure that only one mode of operation is possible there. The transmission lines are assumed to be lossless. At this frequency the connecting lines have electrical length of π, so that the impedance matrix $\boldsymbol{Z}$ is given by

$$\boldsymbol{Z}_{f_1} = \frac{z_1}{4}\begin{bmatrix} 1 & -1 & 1 & -1 \\ -1 & 1 & -1 & 1 \\ 1 & -1 & 1 & -1 \\ -1 & 1 & -1 & 1 \end{bmatrix} \tag{3}$$

where $z_1 = 1/y_1$. The eigenvalues of this matrix are found to be $\lambda_1 = 0$, $\lambda_2 = 0$, $\lambda_3 = 0$ and $\lambda_4 = z_1$. Equation (1) then becomes

$$\begin{aligned} X_1^T \cdot \overline{\mathbf{Z}}i &= 0 \\ X_2^T \cdot \overline{\mathbf{Z}}i &= 0 \\ X_3^T \cdot \overline{\mathbf{Z}}i &= 0 \\ X_4^T \cdot (\lambda_4 I - \overline{\mathbf{Z}})i &= 0 \end{aligned} \tag{4}$$

from which it is concluded that $\overline{\mathbf{Z}}i = kX_4^T$ where k is a scalar constant. If $\overline{\mathbf{Z}}$ is a monotonic function of amplitude then there can be only one solution of k, and the solution is unique.

The eigenvector X_4 at the fundamental frequency (f_1) is given by (5):

$$X_{4f_1} = \begin{bmatrix} 1 \\ -1 \\ 1 \\ -1 \end{bmatrix}. \tag{5}$$

Currents injected into different circuit ports at the fundamental frequency are proportional to the corresponding components of the eigenvector X_{f_1}. Since the impedances seen looking into, all the ports are equal, the voltages at each port are also proportional to the corresponding components of the eigenvector. In order to satisfy the steady state condition for oscillation, the impedance of the device should be equal to the negative of the embedding circuit impedance ($z_{\text{device}} = -z_1$) at all the device ports.

The next step is to calculate the impedance matrix and the eigenvalues of the circuit at the second harmonic. As mentioned above, the impedance matrix of the circuit at the second harmonic has the same general form as the matrix for the fundamental frequency (2). At the second harmonic the active devices are separated by electrical length of 2π (Fig. 3). The impedance matrix for the combining circuit at the second harmonic, looking into the circuit from the device ports is calculated to be

$$\boldsymbol{Z}_{f_2} = \frac{z_2}{4} \begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 \end{bmatrix} \tag{6}$$

where $z_2 = 1/y_2$ is the impedance of the patch antenna, transformed through the feed line at the second harmonic. Just like the fundamental frequency, three of the eigenvalues are equal to zero and the fourth eigenvalue is equal to z_2, so that at the second harmonic only one mode of operation can exist. The eigenvector of the impedance matrix at the second harmonic is given by (7):

$$X_{4f_2} = \begin{bmatrix} 1 \\ 1 \\ 1 \\ 1 \end{bmatrix}. \tag{7}$$

This shows that the currents and voltages at the device ports are equal and in-phase for the second harmonic, therefore the radiated power from the patch antennas is maximum in the broadside direction.

The results obtained so far can be generalized for a periodic circuit containing N active devices. It is clear that the impedance matrix of a circuit of this form containing N devices has only one nonzero eigenvalue at frequencies at which the distance between active elements is an integer multiple of one half wavelength, since this matrix contains only one linearly independent row. This means that at those frequencies, there exists only one mode of operation. At other frequencies where the periodicity of the structure is not an integer multiple of one half wavelength, the determination of the impedance matrix and therefore calculation of circuit's eigenvalues is complicated. Here we calculate the impedance matrix of the circuit at a wide range of frequencies and the circuit's eigenvalues are determined numerically. The plots of real and imaginary parts of eigenvalues for a four diode second harmonic combiner designed to operate at $f_2 = 18$ GHz are shown in Figs. 4 and 5. The components of eigenvectors corresponding to the nonzero eigenvalues at all frequencies (except the ones at which the periodicity of the structure is a multiple integer of one half wavelength), are not equal. This is easy to see, since the circuit impedances looking in from different ports are different except when the periodicity $d = n\lambda/2$. Thus, satisfying the oscillation condition requires the currents through and impedance of the devices to be different.

Under large signal operation, the amplitude of device currents may be different, therefore some undesirable modes of operation can exist. In order to verify the stability of large signal operation at the desirable frequency of operation, large signal simulation of the combining circuit using a relaxation technique [13] was performed. The simulation is a self-consistent harmonic balance solution. The Gunn diodes were represented by a model similar to that given in [14]. A wide range of initial frequencies and amplitudes were used for the relaxation/iteration process; under no circumstances was convergence to other modes of oscillation found, indicating that the mode of operation is unique. The self-locking properties of circuits of this type were also investigated [13]. These simulations showed that the devices remain locked to each other even for a wide range of detuning (at least 10%) and device parameter variation; thus verifying the strong coupling effects that were expected.

Now second harmonic generation by a nonlinear negative resistance device will be considered. When designing the power combining circuit, we need to provide the diodes the impedance z_1 which satisfies the oscillation condition as well as maximum second harmonic generation, and the impedance z_2 should be such that it satisfies the maximum power transfer condition at the second harmonic.

Simulation of the second harmonic generation was performed by Solbach [9]. As the first approximation, it is assumed that the frequency dependence of the negative resistance device can be neglected. The current voltage relationship for the active device is described by a power series:

$$I = a_1V + a_2V^2 + a_3V^3 \tag{8}$$

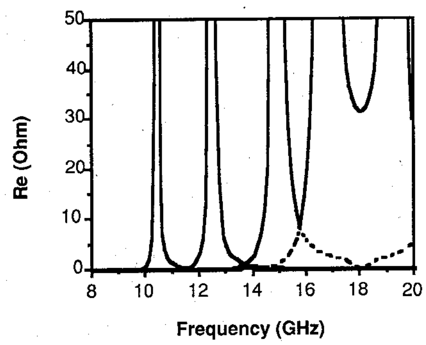

Fig. 4. Plot of real components of eigenvalues for the four diode combiner.

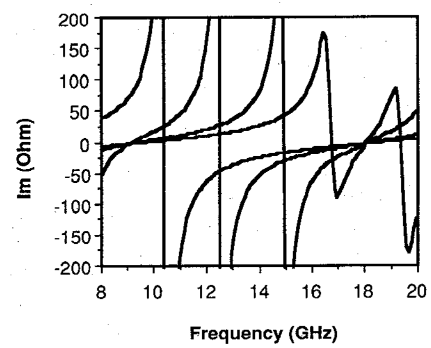

Fig. 5. Plot of imaginary components of eigenvalues for the four diode combiner.

where coefficient a_1 is a negative number and represents the small signal conductance of the active device. It is assumed that the voltage across the active device V, contains only fundamental and second harmonic components and all the higher harmonics are zero.

$$V = v_1 e^{-j\omega t} + v_2 e^{-j\phi} e^{-j2\omega t}. \tag{9}$$

By substituting (9) into the I–V equation for the device and after satisfying nodal equations for the network the second harmonic generation can be analyzed [9]. In an ideal circuit, to maximize the efficiency of second harmonic generation, the resistive part of the impedance seen by the device at the fundamental frequency should be zero. This means no power is dissipated at the fundamental frequency. Also if the load at the second harmonic chosen to be purely real, then analysis shows that the phase difference ϕ between the fundamental and the second harmonic becomes zero. The voltage waveforms across the power combining structure at the fundamental and the second harmonic will then be as shown in Fig. 6, assuming that the diodes' characteristics are the same. This is especially true for monolithic integrated circuits, since all the devices are fabricated on the same chip.

The structure is connected to four microstrip patch antennas to compose an antenna array. The distance between the array elements is equal to the periodicity of the main structure. The patch antennas are resonant at the second harmonic (f_2) and therefore only the second harmonic can be efficiently radiated. As is shown in Fig. 6, all the antennas are in phase. Since the diodes are about a half wavelength apart at the fundamental frequency, the radiated power from discontinuities at f_1 should have its null in the broadside direction, so the interference between the fundamental and the second harmonic is minimal.

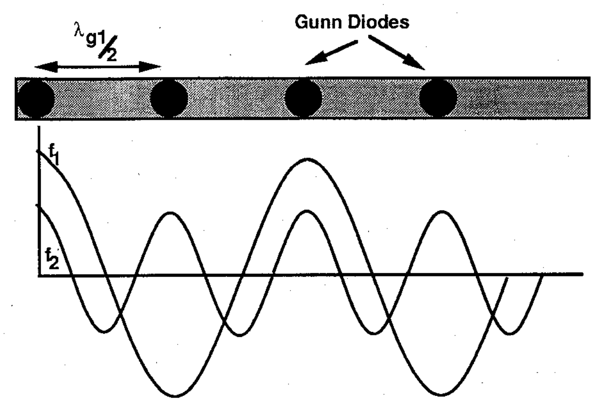

Fig. 6. Voltage waveforms at the fundamental and the second harmonic along the power combining structure.

Design

At the fundamental frequency of oscillation, the microstrip feed lines which are connected to the patch antennas are nearly shorted (since at the fundamental frequency, the antennas are one quarter wave long), therefore they appear as a mostly reactive element shunting the devices. The characteristic impedance of the feed lines can be chosen such that they resonate with the diodes' capacitances at the fundamental frequency. At the second harmonic, the radiation resistance of the patch antennas is transformed by the quarter wave feedlines to an optimum load needed for the diodes to generate the maximum second harmonic. Since the characteristic impedance of the quarter wave feedlines is already determined, the only parameter that can be adjusted is the radiation resistance of the patch antennas. This can be accomplished by altering the width of the microstrip patches. These considerations lead to an initial design which is adjusted as described below.

The next step to design the combiner is to determine the impedance which the negative resistance device should see at the fundamental and the second harmonic in order to generate maximum power at the second harmonic. This is done through a standard technique as follows. The measurement set up is shown in Fig. 7. A single microstrip Gunn diode oscillator is constructed. In this experiment low power packaged Gunn diodes designed to operate at X-band were used. A half-wave microstrip resonator is used to set the desired fundamental frequency of oscillation. The circuit is connected to a spectrum analyzer through a triple stub tuner. By adjusting the stub tuner and trimming the microstrip circuit, the power generated at the second harmonic can be maximized. After removing the Gunn diode, the impedance z_1 of the circuit at the fun-

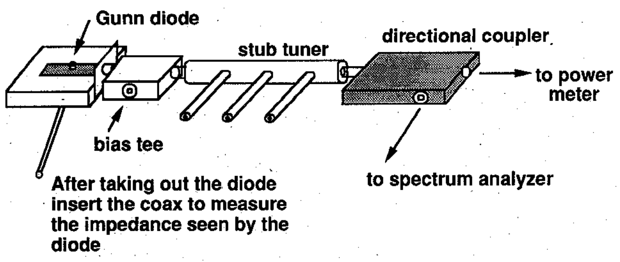

Fig. 7. The measurement set up to optimize the Gunn diode's operation at the second harmonic.

damental and at the second harmonic (z_2) is measured using an automatic network analyzer. The impedances that are measured are those that should be seen by the Gunn diode in order to oscillate at the design fundamental frequency and to generate the maximum second harmonic. Therefore the large signal impedance of the Gunn diode at the fundamental frequency is $-z_1$. Also since z_2 is the impedance to which the Gunn diode delivers maximum second harmonic power, the impedance of the diode at the second harmonic is equal to the complex conjugate of z_2. In our experiment the fundamental frequency at which the Gunn diode generated maximum second harmonic power is 9.3 GHz. The impedances seen by the diode are $z_1 = 5 + j35\ \Omega$ and $z_2 = 20 + j3\ \Omega$. The measurement was repeated on seven similar Gunn devices. The maximum variation of impedances was about 12%. Since the impedance z_1 is mostly reactive, little power at the fundamental is absorbed, and the interaction between the fundamental and the device nonlinearities is maximized, nearly satisfying the conditions described in the previous section for an idealized device. The maximum power obtained from a single Gunn diode oscillator at 18.6 GHz (f_2) was about 5.7 dBm. This information is now used to design the multiple-diode circuit.

In order to provide the diodes the proper impedances (z_1 and z_2), the length and the impedance of the feed line to the antennas, the impedance of the main line, and, slightly, the periodicity of the structure are adjusted so that the impedance looking at each diode port is z_1 and z_2 at f_1 and f_2, respectively. Since the structure is periodic, after providing one of the diodes the correct impedances the others will also see the correct impedances. The circuit design was mainly done using the Touchstone™ microwave circuit analysis program. The patch antennas were designed based on the transmission line model for microstrip antennas [10].

Experiment

A four diode spatial power combining structure was designed and fabricated (Fig. 8), using Rogers Duroid 5880 with a thickness of 20 mil and a relative dielectric constant of 2.2. The antenna radiation pattern is shown in Figs. 9 and 10. For quasi-optical oscillators one may define an effective radiated power (ERP) as the power needed to drive an isotropic radiator to provide the same power density transmitted by the quasi-optical oscillator

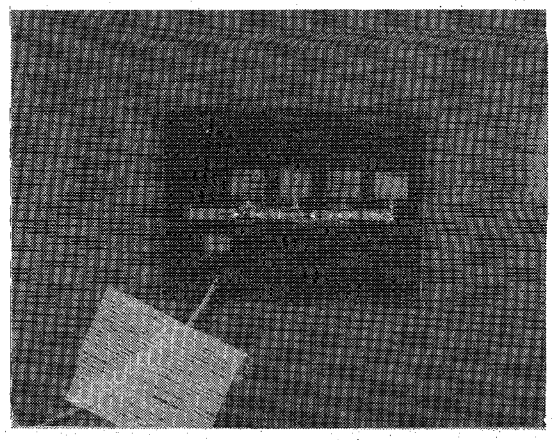

Fig. 8. Photograph of the four diode second harmonic power combining structure.

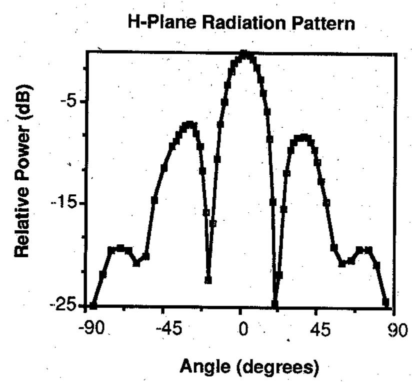

Fig. 9. The *H*-plane radiation pattern of the four diode harmonic combiner.

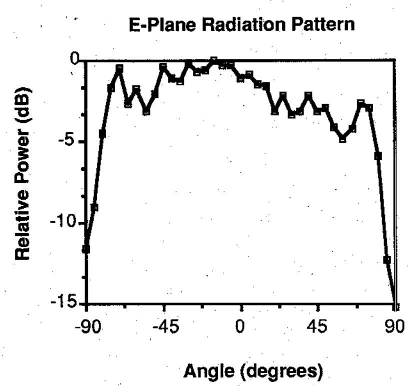

Fig. 10. The *E*-plane radiation pattern of the four diode harmonic combiner.

in a certain direction. The ERP is measured by comparing the output of a standard gain horn driven with a known RF power to that of the quasi-optical oscillator. Also, an isotropic conversion efficiency can be defined as the ratio of the ERP to the dc power consumed by the quasi-optical oscillator [11]. The ERP for the four diode second harmonic power combiner was determined to be 25.7 dBm. The isotropic conversion efficiency was 10.2%. The cross polarization was 17 dB below the peak power in the broadside direction. The power radiated at the fundamental frequency was 15 dB below the power at the second harmonic measured in the broadside direction.

In order to determine the power combining efficiency of this structure the following measurement was performed. The maximum power generated by a single diode oscillator (5.7 dBm) was injected to a single patch antenna. The effective radiated power was measured to be 13.4 dBm. This shows that the ERP from the four diode harmonic combiner (25.7 dBm) is about sixteen times the ERP of a single diode.

Conclusion

A periodic second harmonic spatial power combining oscillator is presented. The oscillators are phase locked at the fundamental frequency. The power is combined in free space. This power combining method makes the use of external resonator circuits unnecessary. This type of circuit structure is compatible with monolithic fabrication. The effective radiated power from a four diode second harmonic combiner is about sixteen times the ERP of a single harmonic generator.

Acknowledgment

We would like to thank Mr. Paul Lamarche of Alpha Industries for providing us the Gunn diodes.

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