Multiple Element Oscillators Utilizing a New Power Combining Technique

Amir Mortazawi, *Member, IEEE,* and Bernard C. De Loach, Jr., *Fellow, IEEE*

Abstract—**A new method is presented to combine the microwave and millimeter wave power generated by many two or three terminal devices. This method forms a single resonant structure from many active devices, and therefore the combiner is stable and does not suffer from simultaneous multimode difficulties. This technique produces compact structures and is readily adaptable to monolithic integration. Through an eigenvalue-eigenvector analysis of the structure, multiple device circuit interactions are investigated. Furthermore, the oscillation stability is demonstrated by large signal analysis of the circuit in the time domain. Experimental results from several power combiners utilizing Gunn diodes are presented.**

Introduction

At millimeter and sub-millimeter wave frequencies individual solid state devices have a limited ability to produce microwave power [1]. In order to obtain higher power, it is very desirable to be able to combine the power generated from many separate devices [2]. The typical method used to combine microwave power from active devices employs some manner of locally resonating the individual active elements. These "tuned" units are then connected together in some one or two dimensional structure capable of adding their powers. Typical configurations place the devices on hybrids [2], at half-wavelength spacings [3], form ring structures with a common load at the center [4], or employ spatial combining techniques [2].

In this paper we present a new combining technique which achieves resonance from the incorporation of the reactive or susceptive elements of the individual gain elements (diodes or transistors) into a filter-like structure. This combiner is fundamentally different from the one presented by Mortazawi and Itoh [3] where individual self resonant oscillators are connected to a transmission line with a periodicity of a half wavelength. An advantage of this technique is that the oscillation is very stable and no simultaneous multimodes are excited. Also due to the fact that the complete structure acts as a single resonator, wideband frequency tuning of such an oscillator is possible.

Analysis of multiple device-circuit interactions as well as results of the large signal analysis of the proposed power combining structure are presented. To demonstrate this combining technique, two, three and four device extended resonance structures utilizing Gunn diodes are designed and fabricated. Also, Rucker's combining technique [4] was employed as a second level combining circuit to construct a four device oscillator from two two-diode resonant structures.

Manuscript received July 7, 1992; revised July 30 1992. This work was supported in part by the National Science Foundation under Grant No. ECS-9110155.

The authors are with the Department of Electrical and Computer Engineering, University of Central Florida, Orlando, FL 32816.

IEEE Log Number 9203709.

Theory

The power combining technique proposed herein is similar to the construction of direct coupled waveguide cavity filters. This technique can be illustrated by considering a two device structure. Fig. 1 displays two negative resistance devices, presumed essentially identical, spaced a distance 1 apart on a transmission line. The distance 1 determines the resonant frequency of the structure for given device susceptances and is chosen such that the susceptive components of the devices cancel. Incorporation of the microwave equivalent circuit for a negative resistance device into the structure of Fig. 1 yields that shown in Fig. 2 where it is assumed that the values of $-G_{1n}$ and jB_{1n} are those appropriate at the desired oscillation frequency normalized to the line's characteristic impedance. The length of the transmission line "l," is chosen such that $-G_{1n} + jB_{1n}$ i.e. the admittance of the first device is transformed to $-G_{1n} - jB_{1n}$ at the end of the line at the operating frequency. By placing another device having the same admittance at this point, the total admittance of the structure becomes $-2G_n$. The terminating load admittance G_L is chosen for maximum power transfer. The similarity of Fig. 2 to that for a single cavity microwave filter suggests that analogues to multiple cavity filters might also be considered as extended resonant structures utilizing multiple devices. For example, Fig. 3 shows a three device combiner as an extension to the above discussion. The susceptance of the middle device, jB_{2n} should be twice the susceptance of the first and the last device ($jB_{2n} = 2jB_{1n}$). This extra susceptance can be provided by connecting a stub to the port of the middle device. In the design of MMIC circuits, the middle device can be fabricated to have twice the capacitance of the first and the last device, obviating the need for the stub. The case of an N device combiner is analogous to that of the three device combiner (the susceptance of all the devices should be twice the susceptance of the first and the last device).

As the number of devices in the combiner increases,

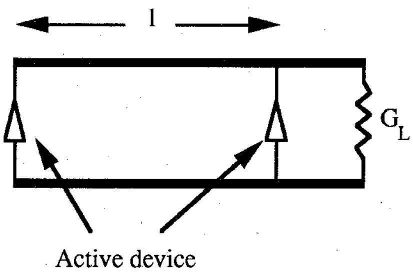

Fig. 1. A two device power combining structure.

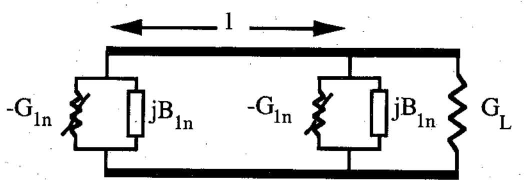

Fig. 2. Equivalent circuit for a two device combiner, length l is chosen such that susceptance of one device resonates with the susceptance of the second device.

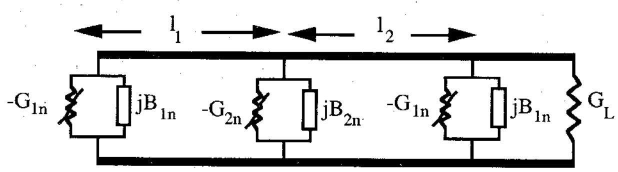

Fig. 3. Equivalent circuit for a three device combiner.

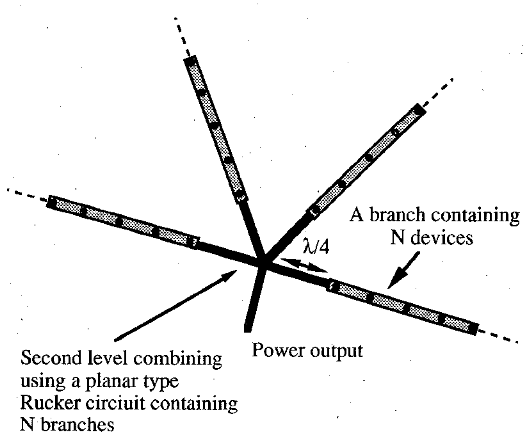

Fig. 4. A method to combine power from many extended resonant structures and to maintain the output impedance at a desirable level.

the net negative resistance of the structure drops. In general this can set a limit to the number of devices that are power combined in a single structure. This is due to the physical dimension of the matching transformer required and circuit losses involved. A method to avoid this limitation is shown in Fig. 4. Assuming the conductance of each device normalized to the line impedance is $-G_n$, the total negative conductance of a branch consisting of N devices is $-NG_n$ when identical devices are employed at all ports (i.e., negative resistance of the branch is 1/N of the negative resistance of a single device). A quarter wave line can then be connected to each branch to transform its conductance to $-1/(NG_n)$. The negative resistance of a single device can be recovered by connecting N branches to each other (i.e., $N(-1/NG_n) = -1/G_n$) using Rucker's method [4] as a second level combiner. This structure (containing N^2 devices) then in turn can be used as a building block to construct a larger combiner. Using this method one can combine the power generated from many devices and simultaneously maintain the output impedance at a desirable level so that the choice of the resistive load for satisfying the oscillation condition will not be a limiting factor.

Analysis

In multiple device oscillators, the potential number of modes of operation is at least equal to the number of devices used in the circuit. There is, however, only one mode of operation that results in maximum power combining efficiency at the design frequency. In order to study the multiple device-circuit interactions and to show that no multiple mode excitations exist, an eigenvalue-eigenvector analysis similar to Kurokawa's method [5] is employed. Based on the oscillation condition, for a multiple device oscillator the following simultaneous equations can be derived:

$$\tilde{X}_n \cdot (\lambda_n I - \overline{Z})i = 0, \; n = 0, 1, 2, 3, N - 1$$

where N is the number of devices, λ_n is the nth eigenvalue of the circuit's impedance matrix looking in from the device ports, x_n is the corresponding eigenvector, I is the identity matrix, $\overline{Z}$ is a diagonal matrix with each of its diagonal elements being equal to the negative of the device impedances and i is a vector containing the currents through circuit ports. For proper power addition it is necessary that all devices operate under the same voltage and current conditions. Therefore all the components of the $\overline{Z}$ should be equal to one of the circuit's eigenvalues for example λ_m. In this case currents injected into all the ports are proportional to the corresponding components of the eigenvector x_m. Since device impedance is a function of the amplitude of the device current, all components of x_m must have the same magnitude, otherwise this will contradict our original requirement that all devices should have the same impedance.

Due to the circuit complexity, numerical techniques are used to determine the eigenvalues and eigenvectors associated with a combiner containing four devices. The results so obtained can be generalized for other values of N. The impedance matrix of a four device combiner at discrete frequencies is calculated using Libra, a commercial microwave circuit analysis program. Then, the eigenvalues and eigenvectors associated with each impedance matrix are determined. The combining circuit is assumed to be lossless and a resistive load is the only lossy element. The device negative resistance is assumed to be $-200\ \Omega$ shunted with a 0.3 pF capacitor lumped in with the combiner circuit for analysis. These values are chosen without having a specific device in mind and only to perform this

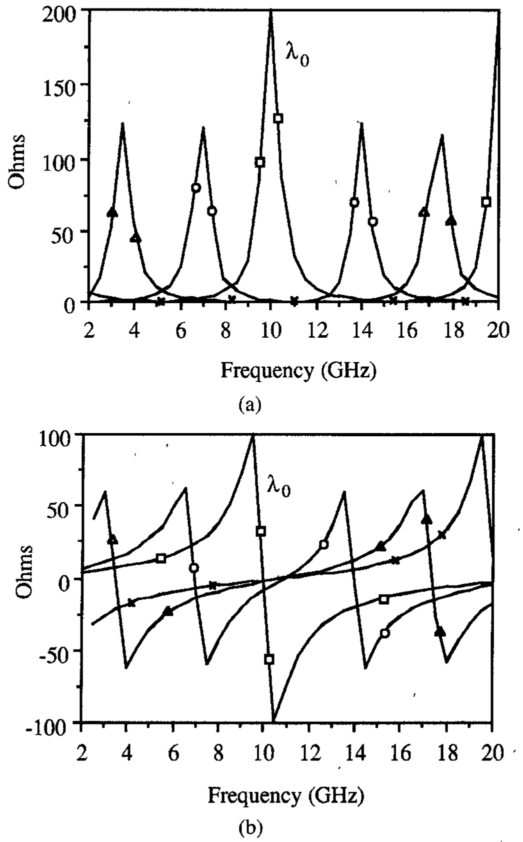

Fig. 5. (a) Plots of real components of eigenvalues for a four device combiner. There are four eigenvalues. The desirable eigenvalue is denoted by λ_0. (b) Plots of imaginary components of eigenvalues for a four device combiner.

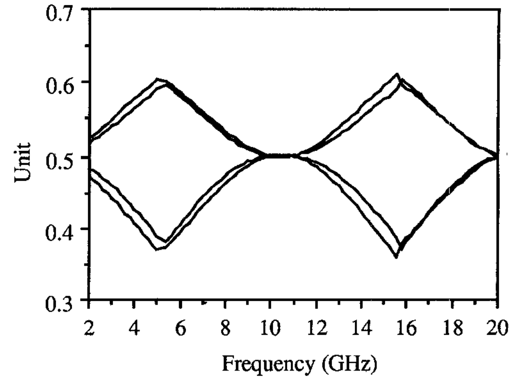

Fig. 6. Plots of magnitudes of eigenvectors associated with the eigenvalue λ_0. These are proportional to magnitudes of currents through four devices.

analysis. This facilitates the interpretation of the results. The characteristic impedance of the lines connecting the four devices is 50 Ω. The plots of the real and imaginary components of the circuit eigenvalues for a four diode combiner are shown in Fig. 5(a) and (b). At the design frequency (10 GHz) only, the negative of the real component of one of the eigenvalues (noted by λ_0) is equal to Z components (200 Ω). The imaginary component of the corresponding eigenvalue intersects zero value at 10 GHz, Fig. 5(b) (since the "device" contains only a negative resistance due to our incorporation of its reactance into the circuit). The components of the other three eigenvalues corresponding to the unwanted modes are very small (compared to λ_0) at 10 GHz and do not satisfy the condition for oscillation, therefore they are not excited. It is then concluded that no simultaneous modes can exist at the design frequency and only a single mode, desirable for power combining operation, is excited. The plot of magnitudes of the four components of the eigenvector (x_0) associated with the desirable eigenvalue (λ_0) is given in Fig. 6. As it can be seen, at 10 GHz all four components have the same magnitude (0.5), which shows that the currents through all four devices should have the same magnitude. The magnitudes of the eigenvectors are also equal at dc and on the order of twice the design frequency. However, in practice, device parasitics make the high frequency resonance unimportant. For those devices that exhibit negative resistance at dc, stabilizing circuits must be employed. There are also other resonances associated with the unwanted modes of operation (for example at 3.47 GHz and 6.49 GHz in Fig. 5(a) and 5(b). By proper choice of the load, which satisfies the maximum power transfer at the desirable mode of operation and the characteristic impedance of the transmission lines, excitation of these extraneous modes can be prevented. An ideal load for the desirable mode makes the circuit appear too lossy for the undesirable modes to be excited (in a parallel resonance configuration, the device negative conductance should be larger than the load's conductance for oscillations to start up).

A large signal analysis of the combiner in the time domain is performed in order to verify oscillator stability and to be able to monitor the operating point for each device. In general the negative resistance of active devices can be modeled using a third order polynomial. This type of approximation is more accurate for Tunnel, Gunn and resonant tunneling diodes whose IV curves can be expressed by a third order polynomial, but can also be used to analyze the behavior of IMPATT and FET based oscillators. Since our goal is to analyze the circuit's performance independent of the type of device utilized, a third order polynomial is used to model a negative resistance device. To further simplify the analysis, it is assumed that the negative resistance is the only nonlinear element. The large signal simulation of a four device combiner is performed utilizing the P-SPICE™ program. The third order polynomial used is given by

$$I = \frac{5}{100} - \frac{1}{200}(V - 10) + 5(V - 10)^3$$

where the small signal negative resistance of the device is −200 Ω. All the devices are biased at 10 volts. The transmission lines are assumed to be lossless. Fig. 7 shows the start up of oscillations at 10 GHz. The voltages corresponding to the first and the second device are denoted as v1 and v2. The voltages across the third and the fourth device are the same as v1 and v2 respectively. The oscillations are initiated from noise and after about 28 nS reach their steady state values. This analysis demonstrates the oscillation stability of the proposed power combining

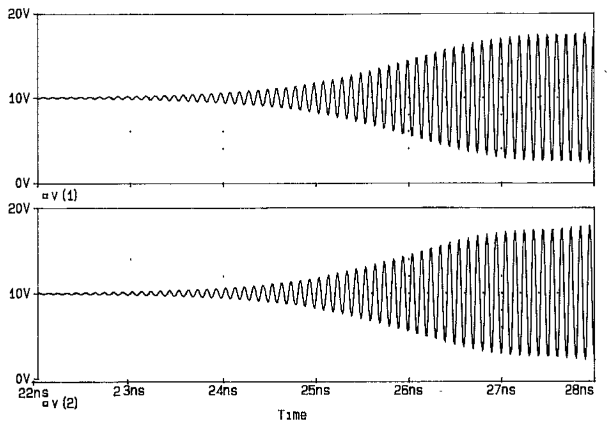

Fig. 7. Plots of start up of oscillations for a four device combiner operating at 10 GHz. v1 and v2 are the voltages across the first and the second device.

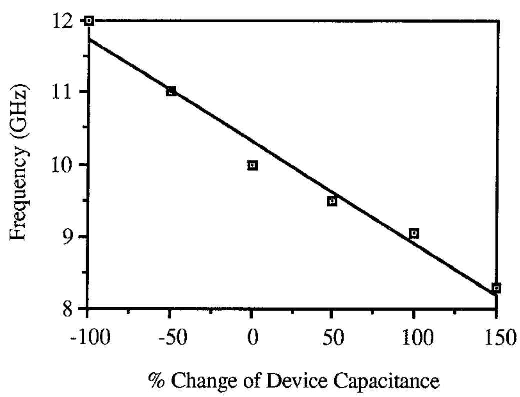

Fig. 8. Plot of frequency of oscillation as a function of devices' capacitance.

structure. The large signal circuit simulation was performed using different device negative resistances and capacitances to account for the variation of device characteristics experienced in the real world. It is observed that the oscillation is not sensitive to device impedance variations up to 30%.

The frequency tuning of this type of power combining structure can potentially be accomplished by shunting each device port with a varactor diode. To simulate the frequency tuning, device capacitances were varied, the result obtained is shown in Fig. 8. A tuning bandwidth in excess of 30% is obtained by varying device capacitances about 200%. The output power variation over the whole frequency band is less than $\pm$ 0.2 dB. In general it is expected that the tuning bandwidth will decrease as a function of number of devices power combined. However, the rate of bandwidth degradation is very dependent on the type of device used, its parasitics, and the variation of its characteristics as a function of frequency. In an ideal case where a negative resistance device is modeled as a nonlinear negative resistance shunted with a nonlinear capacitance, the tuning bandwidth of the combiner is not a function of number of devices power combined. This is due to the fact that the whole structure forms a single resonant circuit.

Experimental Results

Several planar power combining structures were fabricated on Duroid™ substrate with a dielectric constant of 2.3 and thickness of 31 mils. The active elements are X-band low power Gunn devices with dc to rf conversion efficiency of 2% and a bias voltage of 10 V. Several single diode planar oscillators were fabricated and the highest power obtained from any one of them was 13.5 dBm (9.5 GHz). Results from a two, a three and a four device power combiner are presented. To illustrate the structures, Fig. 9(a) and (b) show a two and a three Gunn diode combiner. Biasing is accomplished through a bias tee connected to the output of these structures. This is possible due to the lack of sensitivity to individual device admittances. Table I below shows the output power and frequency of operation of the two, three and four diode combiners.

It should be mentioned that no adjustment was performed to optimize the circuits' performance and the active devices were not matched. The combining efficiencies for three and four diode combiners, compared to the two diode combiner, are 94.3% and 104% respectively. The variation of the oscillation frequency and output power versus bias voltage for these power combiners are shown in Figs. 10 and 11, respectively. Due to the fact that the complete power combining circuit forms a single resonant structure, the circuits are very stable and are not sensitive to bias voltage variation. These structures appear very attractive for the design of high power, frequency tunable sources. The maximum frequency tuning bandwidth for the two, three and four diode combiners are 920 816 and 572 MHz. Phase lock was maintained with bias variations as high as 50% and no discontinuities were observed in frequency or power. As mentioned, in order to maintain the power level while varying the frequency, these structures could be loaded with varactor diodes at each device port.

A Rucker type power combiner [4] for second level power combining was fabricated. The original combining method introduced by Rucker used coaxial lines to combine power from several individual devices. Here a planar Rucker type circuit is investigated which combines power from several branches each containing many devices is investigated. It utilizes two self resonant branches designed based on the power combining technique discussed earlier. Each resonant branch contains two devices shown in Fig. 12. An output power of 18.8 dBm at 9.9 GHz was obtained. The decrease in power compared to the four device combiner mentioned earlier and the shift in the oscillation frequency can be attributed to the variation of device characteristics and circuit elements. It should be pointed out that no circuit adjustment or trimming was

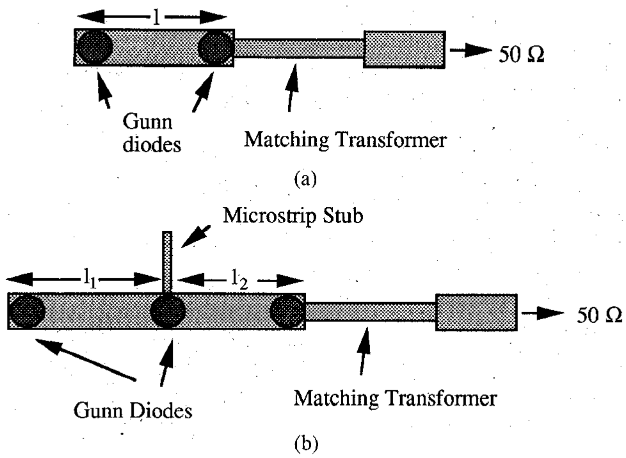

Fig. 9. (a) A two diode microstrip power combiner. (b) A three diode microstrip power combiner.

TABLE I

No. of Devices	Frequency (GHz)	Power (dBm)
Two-diode	9.2	16.9
Three-diode	9.5	19.1
Four-diode	9.1	20.1

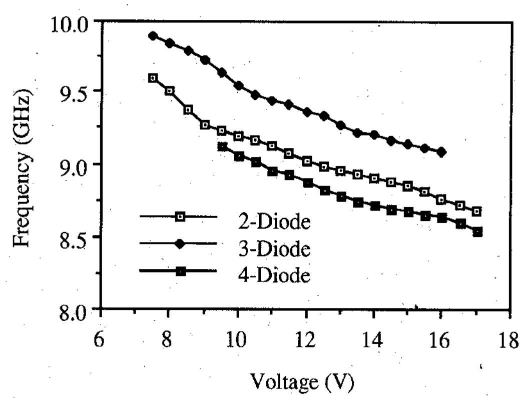

Fig. 10. Plots of oscillation frequency versus bias voltage variation for a two, three and four diode combiner.

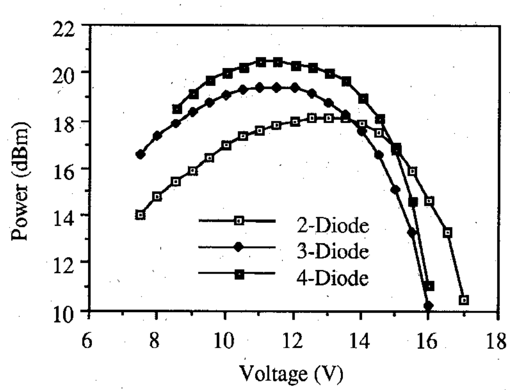

Fig. 11. Plots of output power versus bias voltage variation for a two, three and four diode combiner.

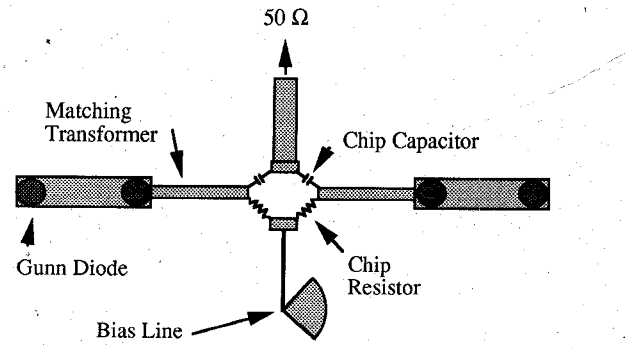

Fig. 12. A four diode combining circuit formed by two two-diode resonant structures utilizing a Rucker type combiner as a second level combining circuit.

performed. This result is important since it demonstrates that the technique presented here can handle a large number of devices by using a Rucker type combiner as a second level combining circuit. In the design of Rucker type planar structures, one obvious limitation imposed on the number of branches that are power combined is due to the coupling between branches and also between branches and the output line as the branches increase in number. Determination of the precise limits on the allowable number of devices requires further investigation.

Conclusion

Based on a new power combining technique which forms a single resonant structure out of many negative resistance devices, several planar power combiners are designed and fabricated. Through an eigenvalue-eigenvector analysis, multiple device-circuit interactions are investigated. To determine the oscillation stability, large signal analysis of the structure is performed. These structures are stable, they do not suffer from simultaneous multimodes and are not sensitive to bias and device parameter variations.

References

[1] B. C. De Loach Jr., *Recent Advances in Solid State Microwave Generators: Advances in Microwaves*, vol. 2, New York, Academic Press, 1967.
[2] K. Chang and C. Sun, "Millimeter-wave power-combining techniques," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 91–107, Feb. 1983.
[3] A. Mortazawi and T. Itoh, "A periodic planar Gunn diode power combining oscillator," *IEEE Trans. Microwave Theory Tech.*, vol. 38, Jan. 1990.
[4] C. T. Rucker, "A multiple-diode high-average power oscillator," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-17, pp. 1156–1158, Dec. 1969.
[5] K. Kurokawa, "The single-cavity multiple-device oscillator," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 793–801, Oct. 1971.

Amir Mortazawi (S'87–M'90) was born in Yazd, Iran in 1962. He received the B.S. degree in electrical engineering from the State University of New York at Stony Brook in 1987 and the M.S. and Ph.D. degrees in electrical engineering from the University of Texas at Austin in 1988 and 1990, respectively.

In 1990, he joined the University of Central Florida as an Assistant Professor. His research interests include millimeter-wave power combining, quasi optical techniques and nonlinear analysis of microwave circuits.

Bernard C. De Loach Jr. (M'57–F'72) received his B.S. and M.S. degrees in Physics from Auburn State University in 1951 and 1952 and his Ph.D. in physics from Ohio State University in 1956. He joined Bell Laboratories in 1956 as a member of the Radio Research Dept. and in 1990 became a Courtesy Professor of Engineering Science at the University of Central Florida. He has worked in the areas of microwave filters, microwave parametric amplifiers, microwave and millimeter wave solid state oscillators, light emitting diodes, semiconductor lasers and avalanche photo detectors.

Dr. De Loach received the IEEE's David Sarnoff award in 1975, and the Franklin Institute's Stuart Ballantine medal also in 1975. He has served in many areas of IEEE activities.