Modeling and Computer Simulation of a Microwave-to-DC Energy Conversion Element

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Abstract—**A microwave-to-dc energy conversion element consisting of a dipole antenna, a low-pass filter, a Schottky-barrier diode, and a dc filter has been modeled using a distributed transmission-line modeling technique that includes skin-effect losses. Computer simulation has shown 80-percent conversion efficiency and has indicated that the diode generates significant power at higher harmonics due to a resonance effect.**

I. INTRODUCTION

THE USE of microwaves for the wireless transmission of electrical energy has been investigated for several years [1]–[8]. More recently it has been proposed to use such a system for the transmission of power from a synchronous orbiting satellite to the earth [9]. Although the system will have a very large transmitting antenna (on the order of 1 km in diameter) and a very narrow transmission beam, the great distances involved will result in a widely dispersed beam on the receiving end of the system. The beam will cover about 100 km^2. The dimensions of the receiving problem led to the development of a distributed, passive, nondirectional microwave-to-dc energy conversion system [8].

The microwave-to-dc energy conversion system uses a large number of identical elements, each of which captures a portion of the transmitted energy and converts the energy to dc. The energy from the individual elements is then combined as dc electrical energy. Thus each of the elements can be considered to act relatively independently except for the problem of matching the entire receiving system to free space in order to prevent the system from reflecting any of the transmitted power. In addition to a dipole receiving antenna, each element contains an input filter to prevent the retransmission of generated harmonics, a Schottky-barrier diode to rectify the microwave energy, and an output filter to smooth the dc output.

Since on the order of ten billion receiving elements will be required in the satellite power station, it is worthwhile to spend a considerable amount of effort to optimize the elements. In order to make some progress in that direction and better understand the operation of the energy conversion element, we have developed a distributed-circuit model of the element and have run computer simulations of its operation.

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II. THE RECTIFYING ANTENNA ELEMENT MODEL

A. The Rectifying Antenna Element

A block diagram of a microwave-to-dc energy conversion element is shown in Fig. 1. The input to each element is a simple dipole antenna. In a receiving system, the individual dipoles are set in a closely spaced array in front of a reflection plane. Thus except for electromagnetic interaction between the dipoles and the ground plane which is used to match the antenna array to free space, the individual elements operate independently in converting the microwave energy to dc.

A short section of transmission line connects the dipole antenna to a five-section stepped-impedance low-pass filter. The filter has a cutoff frequency that is slightly above the transmission frequency. Thus the filter allows power at the transmission frequency to pass from the dipole antenna to the rectifying section but prevents harmonic signals generated by the rectification process from reaching the antenna.

A second longer section of transmission line connects the filter to the output. On this transmission line are a Schottky-barrier diode and a dc output filter capacitor. Although the diode is closer to the filter than the capacitor, the exact position of each on the transmission line is part of the optimization problem.

B. The Transmission-Line Model

Much of the rectifying antenna element is either made up of transmission lines or can be modeled using a transmission line. The section of the element between the dipole and the filter, the filter itself, and the section of the element between the filter and the output are all transmission lines. The dipole antenna requires a transmission line in its model. Thus the modeling of a transmission line will be considered first.

The simplest distributed-transmission-line model is a number of sections of series inductors and parallel capa-

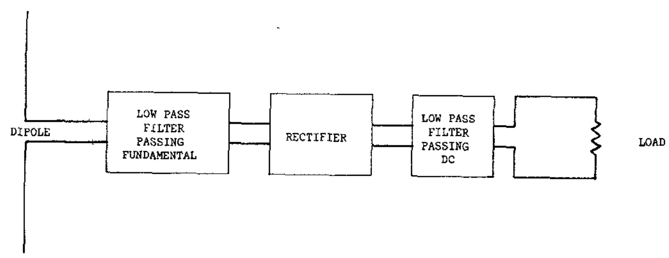

Fig. 1. Dipole-filter-rectifier element.

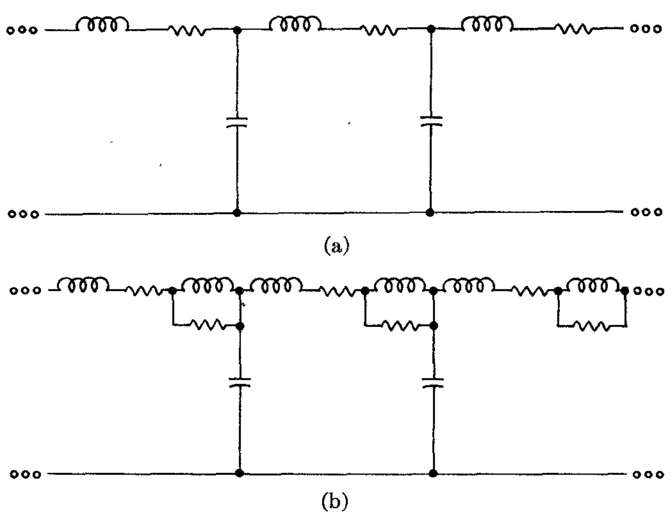

Fig. 2. Distributed-transmission-line models. (a) With resistive losses only. (b) With skin-effect losses.

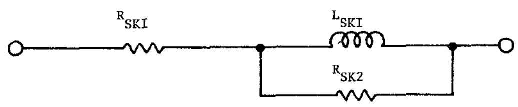

Fig. 3. Circuit model for the skin effect.

citors. Resistive losses in a transmission line can be taken into account by the addition of small resistances in series with the inductors as shown in Fig. 2(a). Skin-effect losses, however, cannot be taken into account so simply since they are a function of frequency. These additional losses should be taken into account to obtain an accurate model since the nonlinearity of the energy conversion generates many significant harmonics.

Skin-effect losses can be taken into account by using the circuit shown in Fig. 3. This circuit has an impedance that increases with increasing frequency and can be used to closely approximate the losses due to the skin effect. With this circuit added in series with each of the inductors in the transmission-line model, a transmission-line circuit model looks like the circuit shown in Fig. 2(b).

C. Dipole Antenna Model

A simple dipole antenna, when used as a receiving element for a single-frequency system, can be modeled as a voltage source in series with an impedance. However, when various frequencies are present, the resonant properties of the antenna must be considered. Thus a model consisting of a voltage source in series with a resistance, a capacitance, and a quarter-wavelength of transmission line was used for the antenna [10]. The capacitance was added principally to prevent the antenna model from acting as a dc short.

A dipole antenna with a length of 6 cm and a width of 2.5 mm was assumed. With these dimensions the quarter-wavelength transmission line in the model has an impedance of 500 Ω while the series resistance has an impedance of 2500 Ω. The series capacitance was set to 1 pF which results in an added impedance which is small when compared to the 2500-Ω series resistance.

The input power to the element is the power input to the capacitor in the antenna model. Thus the power loss in the input resistance is not considered as a loss in the rectifying antenna system but only part of the required matching process.

D. Low-Pass Input Filter

Since detailed data on the input filter from an actual rectifying antenna were not available, the filter section for the model had to be designed A five-section stepped-impedance Chebyshev prototype filter was chosen with a cutoff frequency of 2.64 GHz to obtain minimal attenuation at the fundamental frequency of 2 5 GHz [11]. This transmission-line filter design also resulted in a bandpass from 14.9 to 20.2 GHz.

The filter was modeled using fifteen transmission-line model sections, three for each section of the filter. The characteristics of the filter were confirmed by using the Electronic Circuit Analysis Program (ECAP) to calculate its response.

E. Schottky-Barrier Diode Model

A real Schottky-barrier diode cannot be modeled simply as a rectifying junction at microwave frequencies. The details of the junction as well as the lead inductance and case capacitance all must be taken into account. A circuit model that does so is shown in Fig. 4. The model includes an ideal Schottky metal-semiconductor junction characteristic, the varying junction capacitance C_j, whose value is a function of the junction voltage; the varying high-resistivity layer resistance R_{su}, whose value changes linearly with the junction depletion depth; the fixed resistances due to the diode substrate R_{ss}; ohmic contact R_{sc}; and lead resistance R_{sl}; the lead inductance L_D; and the case capacitance C_p.

Initially, the total fixed resistance $R_{\text{FIX}} = R_{sl} + R_{sc} + R_{ss}$, the lead inductance L_D, and the case capacitance C_p, were set to 1 Ω, 1 nH, and 0.2 pF, respectively, as param-

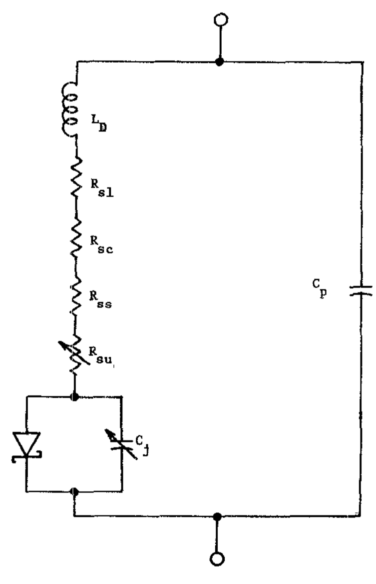

Fig. 4. Schottky-barrier diode model: C_p, package capacitance; L_D, wire lead inductance; R_{sl}, wire lead series resistance; R_{sc}, contact series resistance; R_{ss}, substrate series resistance; R_{su} undepleted region series resistance; C_j, junction capacitance.

eter reference points. The other parameters in the model—the junction saturation current, the high-resistivity layer resistance, and the junction capacitance—are actually functions of other, more fundamental, diode parameters.

The semiconductor material was initially assumed to be gallium arsenide and the metal to be either platinum or gold resulting in a junction barrier height of 0.9 eV. A junction area of 10^{-4} cm^2, a high-resistivity layer depth of 10 μm, an n-type impurity concentration of 5×10^{15} donors/cm^2, and a temperature of 300 K were also taken as initial parameter values.

F. Overall Model Description

The complete circuit model of a rectifying antenna element is shown in Fig. 5. The input to the circuit consists of the voltage source, resistance, capacitance and ten T-type transmission-line model elements forming the dipole antenna model. The next two transmission-line model sections form the short transmission line between the antenna and the input filter. The filter is modeled by the next 15 transmission-line model sections. The remaining 20 transmission-line model sections are associated with the transmission line between the filter and the output. The diode model is attached to one of these transmission-line model sections as can be seen in the figure. Initially, the diode was located at the thirty-sixth section. The output filter capacitor was added to the circuit by simply increasing the value of one of the parallel capacitances. A value of 100 pF at the forty-sixth section was used for this capacitance. The final element in the model is the load resistance with an initial parameter value of 100 Ω. Thus, as can be seen, the major component in the model consists of the T-type transmission-line model elements.

Besides the aforementioned diode parameters, the parameters in the rectifying antenna element that can readily be studied using the above model are the diode and dc filter position and the load impedance.

III. SIMULATION OF RECTIFYING ANTENNA ELEMENT

A. Equations Describing the Operation of the Element

The complete model circuit shown in Fig. 5 can be described using 146 first-order differential equations and several algebraic equations. The first three node voltages V_A, V_{S1}, and V_{S2} are computed using the following three algebraic equations:

$$V_A = V_{A_{\max}} \sin(2\pi f t) \tag{1}$$

where f is the operating frequency of 2.5 GHz and $V_{A_{\max}}$ is

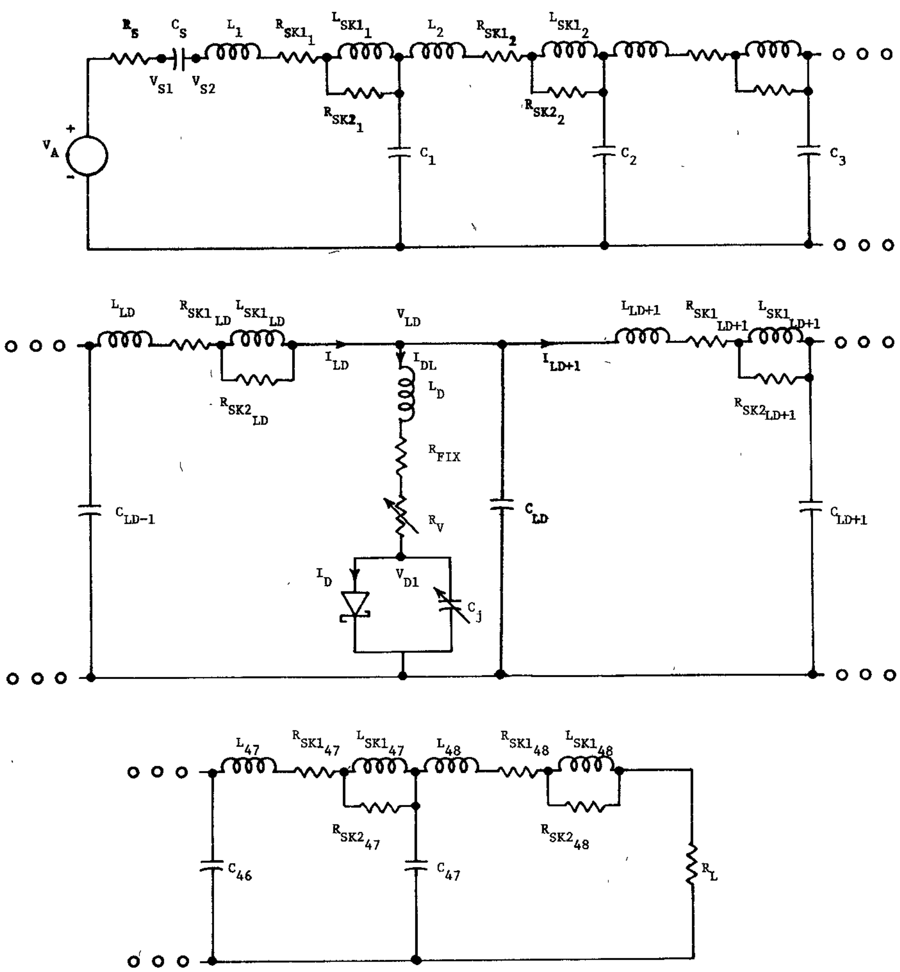

Fig. 5. Complete model of rectifying antenna element.

the peak input voltage

$$V_{S1} = V_A - R_S I_1 \tag{2}$$

where R_S is the input matching resistance and I_1 is the input current, and

$$V_{S2} = V_{S1} - V_{CS} \tag{3}$$

where V_{CS} is the voltage across the input capacitance. The voltage across the input capacitance is controlled by the differential equation

$$\frac{dV_{CS}}{dt} = \frac{I_1}{C_S} \tag{4}$$

where C_S is the input capacitance.

The bulk of the circuit is described by a set of equations which calculate the voltages and currents on the distributed-transmission-line model

$$\frac{dI_j}{dt} = \frac{V_{j-1} - R_{SK1_j} I_j - R_{SK2_j}(I_j - I_{SK1_j}) - V_j}{L_j}, \quad j = 1,48 \tag{5}$$

$$\frac{dI_{SK1_j}}{dt} = \frac{(I_j - I_{SK1_j}) R_{SK2_j}}{L_{SK1_j}}, \quad j = 1,48 \tag{6}$$

and

$$\frac{dV_j}{dt} = \frac{I_j - I_{j+1}}{C_j}, \quad j = 1,47. \tag{7}$$

In the foregoing equations, R_{SK1_j} and R_{SK2_j} are resistances in a skin-effect circuit, L_{SK1_j} is an inductance in a skin-effect circuit, L_j is a series transmission-line inductance, C_j is a parallel transmission-line capacitance, I_{SK1_j} is a current in a skin-effect circuit inductor, I_j is a transmission-line series current, and V_j is a transmission-line voltage. Also, $V_0 = V_{S2}$ is one of the input voltages and $V_{48} = V_L$ is the load voltage.

The previous equations do not take the diode into account. The diode current modifies one of the parts of (7) so that

$$\frac{dV_{LD}}{dt} = \frac{I_{LD} - I_{LD+1} - I_{DL}}{C_{LD}} \tag{8}$$

where LD is the location of the diode on the transmission line and I_{DL} is the diode lead current. (The diode case capacitance is taken into account by increasing the transmission-line parallel capacitance at the location of the diode.)

The diode junction voltage can be found from the stored charge in the junction

$$V_{D1} = V_{bi} - \frac{Q_D^2}{2\epsilon q N_D A^2} \tag{9}$$

where V_{D1} is the junction voltage, V_{bi} is the built-in junction voltage, Q_D is the stored charge in the junction, ϵ is the semiconductor permittivity, q is the electron charge, N_D is the semiconductor donor concentration, and A is the junction area. The junction current is related to the junction voltage by the diode equation

$$I_D = I_S \left[\exp\left(\frac{qV_{D1}}{kT}\right) - 1\right] \tag{10}$$

where I_S is the junction saturation current, k is Boltzmann's constant, and T is absolute temperature. The resistance of the high-resistivity semiconductor region under the junction is a function of the junction voltage

$$R_V = \frac{1}{Aq\mu N_D}\left[d_{ND} - \left(\frac{2\epsilon}{qN_D}(V_{bi} - V_{D1})\right)^{1/2}\right] \tag{11}$$

where μ is the semiconductor mobility and d_{ND} is the depth of this high-resistivity region. The rate of change of the lead current is a function the voltage across the lead inductance

$$\frac{dI_{DL}}{dt} = \frac{V_{LD} - I_{DL}(R_{FIX} + R_V) - V_{D1}}{L_D} \tag{12}$$

where R_{FIX} is the sum of resistances of the lead wire, ohmic contact, and semiconductor substrate, and L_D is the lead inductance. The rate of change in the junction charge is computed from the difference between the junction current and lead current

$$\frac{dQ_D}{dt} = I_{DL} - I_D. \tag{13}$$

Finally, the load voltage is found from

$$V_L = R_L I_{48} \tag{14}$$

where R_L is the load resistance and I_{48} is the current in the last of the transmission-line model sections.

B. Computer Program Description

The previous equations were programmed for parallel integration using a predictor–corrector integration method on an IBM 370/158. The program was allowed to run for a sufficient time to allow the startup transients to die down and achieve steady-state operation. The achievement of steady-state operation was measured by checking the cycle-to-cycle change of several waveforms on the model. A relative root-mean-square change of about 10^{-5} was readily achievable.

The program has several types of optionally available outputs: a listing of the calculated values of a number of variables, line-printer and Calcomp plots of several variables, fast Fourier transform of various variable waveforms to obtain frequency spectrums, the power transfer at various points in the circuit, the conversion efficiency calculated from the average output power divided by the input power, and finally the cycle-to-cycle change error estimates.

C. Some Preliminary Results

The simulation program was run until steady state was achieved with an input voltage of 300 V_{rms} and with the

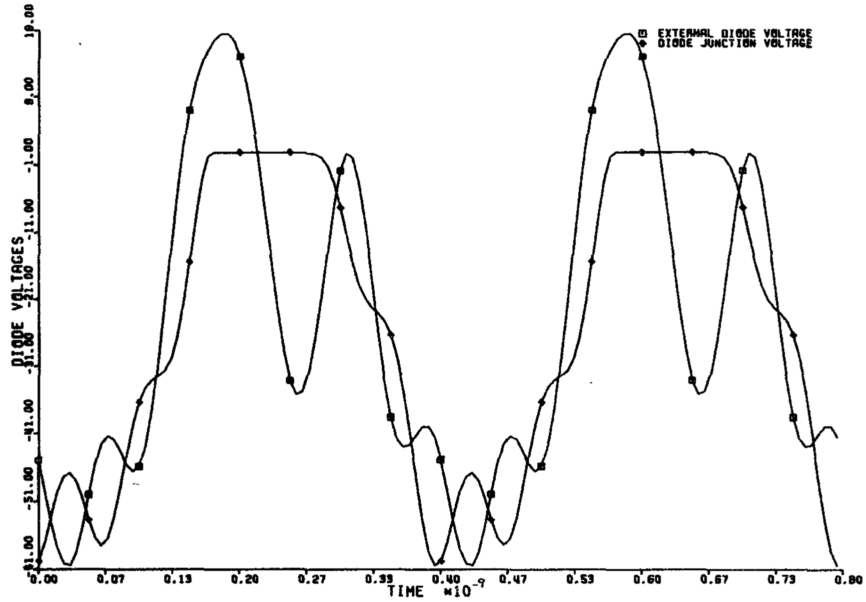

Fig. 6. Diode voltages versus time.

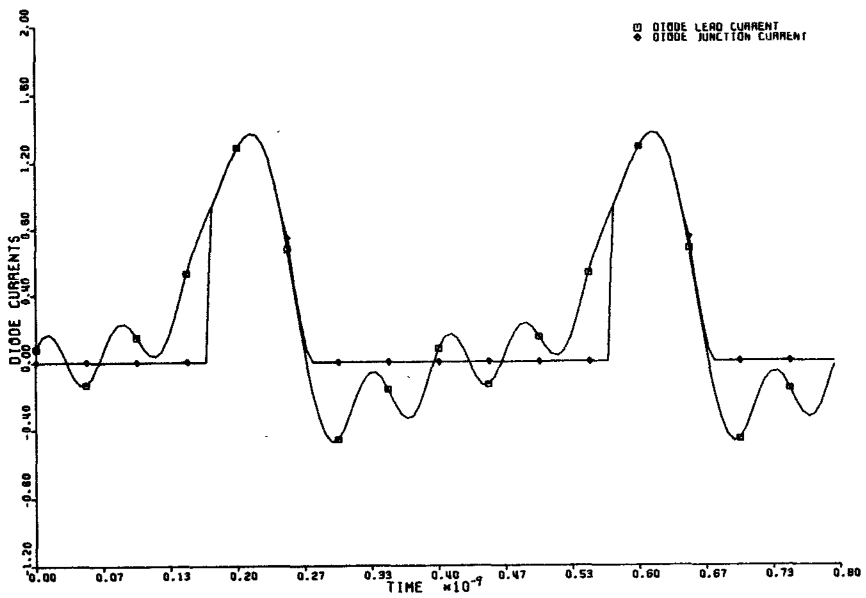

Fig. 7. Diode currents versus time.

previously described reference parameters. With this voltage, the system had an input power of 7.25 W and a dc output power of 5.80 W resulting in an efficiency of 80 percent. Of the 1.45 W of lost power, 1.08 W was lost in the diode and 0.37 W in the transmission lines. Surprisingly, most of the loss in the diode is in the series resistance (0.89 W) rather than in the junction (0.19 W).

Fig. 6 shows both the diode junction voltage and the external diode voltage as functions of time, while Fig. 7 shows the diode junction current and the diode lead current as functions of time. As expected, the voltage waveforms are not simple rectified sinusoids but rather complicated functions containing many harmonics. In fact, the presence of a rather strong resonance is easily noticed in those portions of the waveform when the diode is reverse-biased.

As shown in Fig. 8, the diode junction capacitance, lead inductance, and case capacitance and transmission-line capacitance form a resonant loop. When the diode is reverse biased this loop rings since it is very underdamped. With the diode turned off, the loop has a resonant frequency of about 13 GHz and a Q of about 30. A plot of the approximate resonance frequency as determined from time plots of the lead current of the type shown in Fig. 7 versus $(L_D)^{1/2}$ is shown in Fig. 9. As can be seen, the points form a straight line confirming that L_D is the primary inductive component in the resonant circuit.

The resonance explains the large loss in the diode series resistances. When the diode is reverse biased, there is very little junction power loss. However, the loop current, due to the resonance, results in losses in the series resistances even though the diode is "turned off."

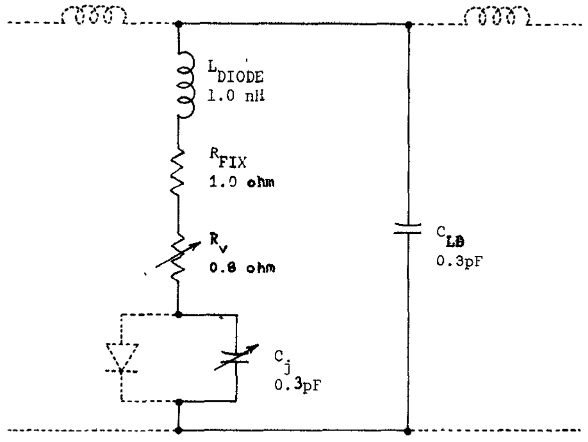

Fig. 8. Diode resonant loop under reverse bias.

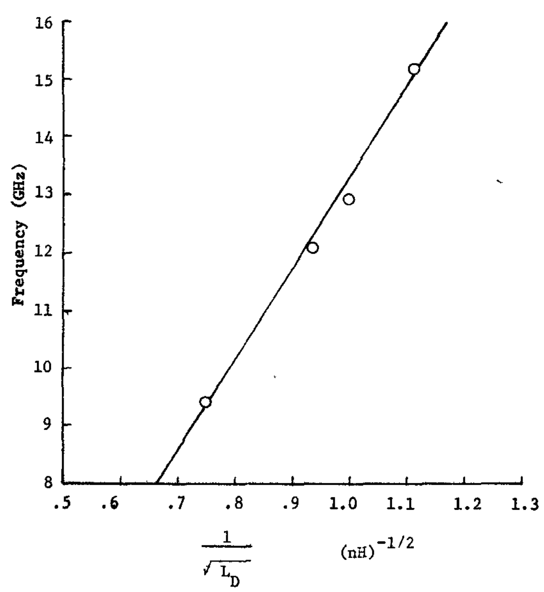

Fig. 9. Diode resonant frequency as a function of L_D.

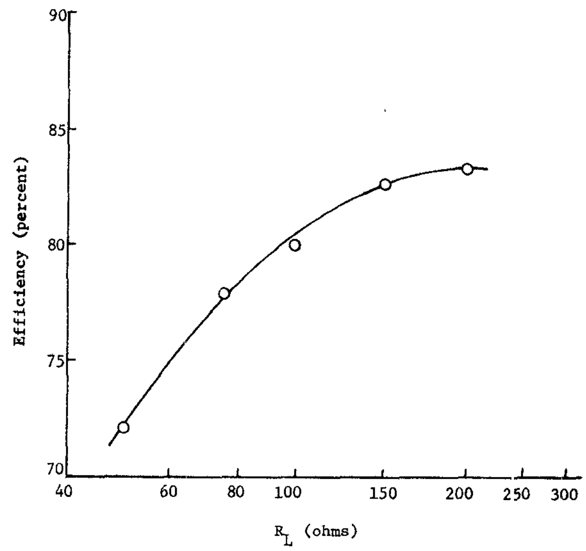

Fig. 10. Conversion efficiency as a function of load resistance.

A preliminary parameter study of efficiency versus load resistance is shown in Fig. 10. As can be seen, there is only a very weak dependence of the efficiency on this parameter.

IV. CONCLUSION

It has been shown that it is possible to model a nonlinear microwave system, in particular a microwave-to-dc energy conversion element, using a distributed-transmission-line modeling technique that includes skin-effect losses. This model and simulation technique is currently being used in a optimization study of the efficiency of a microwave-to-dc energy conversion element as a function of the diode parameters mentioned in Section III-E, the relative diode and filter positions, and the element load. Currently, a computer simulation using the model has shown an efficiency of 80 percent with most of the energy loss coming in the diode rectifier as expected. The simulation has also shown that there is a series resonance effect in the Schottky-barrier diode that tends to generate strong higher harmonics and results in significant power losses when the diode is reverse biased.

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