# Theoretical and Experimental Development of 10 and 35 GHz Rectennas

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*Abstract*—**A 35 GHz rectenna has been developed with 39% conversion efficiency. The rectenna uses a microstrip dipole antenna and a commercially available mixer diode. Over 60% conversion efficiency was demonstrated using this diode at 10 GHz. A theoretical analysis was derived to predict the performance of the rectenna. The analysis is a useful tool for device and circuit design. The theoretical and experimental results should have many applications in microwave power transmission and detection.**

## I. INTRODUCTION

THE AVAILABILITY of power for use in space is a key requirement for future space activities. The current method of power generation with solar panels or batteries on board of each satellite has many difficulties in packaging the power system, unfolding them in space, and achieving a high output power. Power becomes a limiting factor in the design of most systems. To overcome these problems, a potential method may be to deploy a Utility Power Satellite (UPS) in space. The UPS will generate power using solar, nuclear or other techniques. The power would then be converted into microwave or laser beams, transmitted through a free space and converted back into a useful form of energy for users.

The laser beam has an advantage of the small beam divergency. However, the efficiencies in generating the laser beam and converting it back into electrical energy are low compared with those of microwave. Therefore, the development of power transmission system by microwave beams is attracting new attention in space application.

The rectenna (rectifier + antenna) which receives and converts microwave power into dc power is a key element of this power transmission system. Since the rectenna was invented [1] it has been improved through studies [2], [3] and used for various applications such as the microwave powered helicopter [4], the receiving array for Solar Power Satellite [5], and the experiment on the microwave powered aircraft which was recently conducted by Canada under the project name of SHARP (Stationary High Altitude Relay Platform) [6]. As a result, the structure of rectenna has been evolved from a bulky bar-type to a planar thin-film type greatly reducing the weight to power output ratio [7]. The typical power conversion efficiency achieved is 85% at 2.45 GHz.

The rectifying process is a nonlinear process. It is difficult to figure out how the rectenna circuit is optimized for the maximum conversion efficiency. There were several theoretical analyses to solve this problem. These analyses can be divided into two methods. One is to directly simulate the rectenna circuit in time domain [8]. The other is to find a closed-form equation which can explain the relationship between diode parameters and the conversion efficiency [9], [10]. A computer simulation in [8] sucessfully showed the conversion efficiency lager than 80% at 2.5 GHz. A mathematical model in [9] was derived to relate the power conversion efficiency with the conduction period of diode. A fixed relationship between the dc output voltage and the RF input voltage was assumed and the conduction period was unknown. A closed-form for efficiency in [10] was derived by expanding the current-voltage ($I$–$V$) equation of diode with the diode voltage. But this is valid only for a high-efficiency rectenna since short turn-on period was assumed.

All these studies were concentrated on 2.45 GHz. Considering that the frequencies below 3 GHz are not strongly attenuated by the atmosphere even under a severe weather condition [11], 2.45 GHz is thought of as a proper frequency for the application of power transmission between ground-to-ground, ground-to-space, and space-to-ground. However, for the space-to-space application the operating frequency can be increased to allow power transmission over much longer distances with the smaller antenna and rectenna. Although the efficiencies of rectenna and generator are low at 35 GHz, the advantages of size reduction and longer transmission distance overcome the low device conversion efficiencies. The overall system efficiency is thus higher at 35 GHz than 2.45 GHz for long distance transmission using the same size of antenna and rectenna [12].

In this paper we report theoretical analysis and experimental results on 10 and 35 GHz rectenna. A closed-form equation for the conversion efficiency has been derived to analyze the diode for the high frequency rectenna. To derive the closed-form equation we assumed that the effects of harmonics higher than or equal to the second order were small and the forward voltage drop of the diode didn't change during the turn-on period. The closed-form equation was validated by comparing it with a nonlinear circuit

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simulation computer program called LIBRA developed by EEsof Inc.. The maximum efficiency limited by the parasitic series resistance and the junction capacitance of a diode has been calculated using the closed-form equation. It was possible to interpret the cutoff frequency of a diode as a parameter to limit the highest operating frequency of the diode in the power conversion circuit as a result of this calculation. This result of closed-form equation will be useful to select or design a diode for rectenna at any arbitrary frequency.

A 35 GHz dipole rectenna has been designed with the closed-form equation and LIBRA which uses a Harmonic Balance method to solve a nonlinear circuit. The overall conversion efficiency of rectenna was measured using the waveguide array simulator. The dc power was 39% of the input power of 51 mW and the reflected power was 12% of the incident power. The incident power includes both the input power to the rectenna diodes and the reflected power. The power conversion efficiency of the diode at 10 GHz was measured by sending the power through a coaxial line. The highest efficiency was measured as 60%, which was consistent with the simulation result with LIBRA. The devices used were commerically available mixer diodes which are not optimized for this application. It is believed that higher efficiency could be achieved with optimized devices.

## II. Closed-Form Equation for the Conversion Efficiency

### A. Basic Circuit Structure of Rectenna

The basic structure of a rectenna is shown in Fig. 1. The low pass filter inserted between the antenna and rectifying circuit is designed so that the fundamental frequency can be passed and a significant portion of the higher order harmonics generated from the nonlinear rectifying circuit be rejected back to the rectifying circuit. The rectifying circuit consists of a single diode shunt-connected across the transmission lines.

The basic principle of the microwave power conversion by ths rectifying circuit is analogous to a diode clamping circuit or a large signal mixer at microwave frequencies. The power conversion efficiency is maximized by substantially confining all the higher order harmonics between the low pass filter and the dc pass filter, using an efficient diode and matching the diode's input impedance to antenna's input impedance.

The power conversion efficiency of a diode changes as the operating power level changes. The power conversion efficiency of a rectenna generally changes with the input power as shown in Fig. 2. $V_J$, $V_{br}$, and $R_L$ are the forward voltage drop (junction voltage), the breakdown voltage of the diode, and the dc load resistance of the rectenna, respectively. The efficiency is small in the low power region because the voltage swing at the diode is below or comparable with the forward voltage drop of the diode. The efficiency increases as the power increases and levels off with the generation of strong higher order harmonics. The efficiency sharply decreases as the voltage swing at the diode exceeds the breakdown voltage ($V_{br}$) of the diode. The critical input power where the breakdown effect becomes dominant is expressed as $V_{br}^2/4R_L$ in Fig. 2.

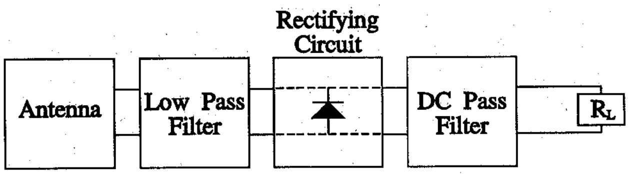


Fig. 1. Block diagram of a rectenna circuit.

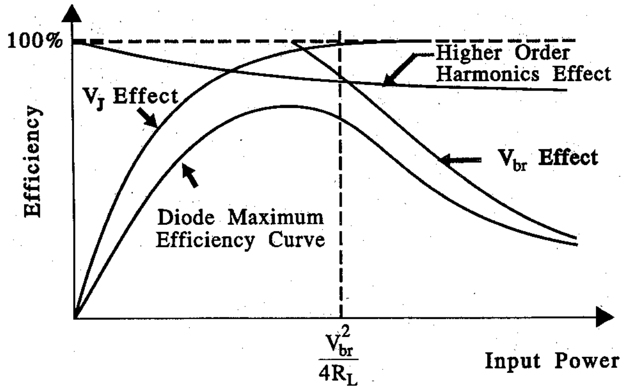


Fig. 2. General relationship between microwave to dc power conversion efficiency and input power.

The three parameters of diode such as $V_{br}$, zero-biased junction capacitance ($C_{jo}$), and series resistance ($R_S$) determine the power conversion efficiency of rectenna. These parameters are related with each other due to the diode material properties [13]. The breakdown voltage of a diode used for a 2.45 GHz rectenna is on the order of 60 V. The $C_{jo}$ and $R_S$ with this breakdown voltage are several pF and less than 1 Ω respectively which are small enough to operate at 2.45 GHz. The breakdown voltage is much larger than the junction voltage of the diode which is about 0.8 V. Therefore the efficiency of the efficiency vs input power curve in Fig. 2 can increase nearly to 100% before reaching at the region where the breakdown effect is dominant. For 35 GHz rectenna, $C_{jo}$ should be reduced to the order of 0.1 pF. But this reduction results in the increase of $R_S$ and the decrease of $V_{br}$. The typical value of $V_{br}$ of a *Ka*-band mixer diode is 10 V with $R_S$ of 4–8 Ω and $C_{jo}$ of 0.1 pF. The efficiency would be low with this small value of $V_{br}$ because the junction voltage effect is still not negligible even at the maximum efficiency point of the efficiency vs input power curve as illustrated in Fig. 2. Consequently, it is important to have a good trade-off between the diode parameters in designing a diode. The closed-form equation derived in the next subsection will be useful to design a diode for a high frequency rectenna.

### B. Derivation of Closed-form Equation for Conversion Efficiency

The equivalent circuit of the Schottky diode is shown in Fig. 3. The diode consists of a series resistor ($R_S$), a

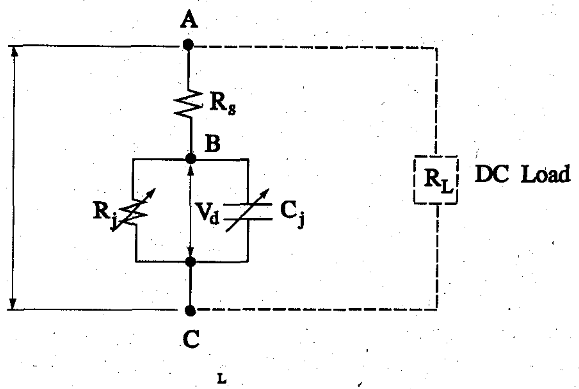


Fig. 3. Equivalent circuit of a *Ka*-band beamlead Schottky diode used for the derivation of a closed-form equation. $R_j$ and $C_j$ model the intrinsic junction for the diode, $R_S$ is the parasitic series resistance of the diode and $R_L$ is the dc load.

nonlinear resistor ($R_j$) described by the $I$–$V$ relationship of a diode at dc, and a nonlinear junction capacitance ($C_j$). The reactive parasitic elements of the physical diode are excluded from this model because they can be tuned out without affecting the rectifying efficiency. The closed-form expressions for the rectifying efficiency and the effective microwave impedance of the diode are derived with the following assumptions:

1) The voltage $V$ applied between node $A$ and $C$ consists of the dc term and the fundamental frequency only.
2) The current due to the junction capacitance is negligible when the diode is forward biased.
3) The forward voltage drop across the intrinsic diode junction is constant during the forward-bias period.

The first assumption is reasonable considering that the magnitude of the higher order harmonics is usually small compared to those of the dc and fundamental components. The second assumption is based on the fact that the change of $V_d$ in Fig. 3 is small during the forward-bias period.

Under these assumptions, the voltage waveform of $V$ and $V_d$ can be expressed as follows:

$$V = -V_0 + V_1 \cos(\omega t) \tag{1}$$

$$V_d = \begin{cases} -V_{d0} + V_{d1} \cos(\omega t - \phi), & \text{if diode is off} \\ V_f & \text{if diode is on.} \end{cases} \tag{2}$$

The voltage $V_0$ in (1) is the output dc voltage and the voltage $V_1$ is the peak voltage of an incident microwave. $V_{d0}$ and $V_{d1}$ are the dc and the fundamental frequency components of diode junction voltage $V_d$, respectively, when the diode is off. $V_f$ is the forward voltage drop of the diode when the diode is on. $V_0$ is known by specifying the output dc power. The efficiency and the diode effective impedance are calculated by relating the other variables such as $V_1$, $V_{d0}$, $V_{d1}$, and $\phi$ to the known variable of $V_0$.

By applying Kirchoff's voltage law along the dc-pass loop shown as a dotted line in Fig. 3, the dc voltage ($-V_{d,\mathrm{DC}}$) of $V_d$ is related to the dc component of $V$ according to

$$V_0 = \frac{V_{d,\mathrm{DC}}}{1 + r} \tag{3}$$

where $r = R_s/R_L$ and $R_L$ is the dc load resitance as shown in Fig. 3. $V_{d,\mathrm{DC}}$ introduced here is the average value of the waveform $V_d$, and $V_{d0}$ in (2) is the dc component of the waveform $V_d$ in turn-off period. Therefore, $V_{d,\mathrm{DC}}$ is derived by taking an average of $V_d$ over one full period as

$$V_{d,\mathrm{DC}} = V_{d0}\left(1 - \frac{\theta_{\mathrm{off}}}{\pi}\right) + \frac{V_{d1}}{\pi} \sin \theta_{\mathrm{off}} - V_f \frac{\theta_{\mathrm{off}}}{\pi} \tag{4}$$

where $\theta_{\mathrm{off}}$ is the phase angle where the diode is turned off as shown in Fig. 4. The phase variable $\theta$ is defined as $\theta = \omega t - \phi$. Since the switching of the diode occurs when $V_d$ is equal to the forward voltage drop $V_f$ of the diode, $\theta_{\mathrm{off}}$ is calculated by

$$\cos \theta_{\mathrm{off}} = \frac{V_f + V_{d0}}{V_{d1}}. \tag{5}$$

On the other hand, the equation for the current flowing through $R_S$ is written as follows when the diode is off:

$$R_S \frac{d(C_j V_d)}{dt} = V - V_d. \tag{6}$$

Since $C_j$ is a monotonic increasing funtion of $V_d$, $C_j$ can be expanded as follows:

$$C_j = C_0 + C_1 \cos(\omega t - \phi) + C_2 \cos(2\omega t - 2\phi) + \cdots. \tag{7}$$

Substituting $C_j$ into (6) with a Fourier series of (7) and neglecting the terms higher than the second harmonic, the equation becomes

$$\begin{aligned} \omega R_S(C_1 V_{d0} - C_0 V_{d1}) \sin(\omega t - \phi) &= V_{d0} - V_0 + (V_1 \cos \phi - V_{d1}) \cos(\omega t - \phi) \\ &\quad - V_1 \sin \phi \sin(\omega t - \phi). \end{aligned} \tag{8}$$

Since the above equation should hold during the off period of the diode, each term should separately zero:

$$V_{d0} = V_o \tag{9a}$$

$$V_{d1} = V_1 \cos \phi \tag{9b}$$

$$V_1 \sin \phi = \omega R_S(C_0 V_{d1} - C_1 V_{d0} + C_2 V_{d1}). \tag{9c}$$

The phase delay is obtained from (9) as follows:

$$\begin{aligned} \phi &= \arctan\left[\omega R_S \left(C_0 - \frac{C_1 \cos \theta_{\mathrm{off}}}{1 + v_f} + C_2\right)\right] \\ &= \arctan(\omega R_S C_{\mathrm{eff}}) \end{aligned} \tag{10}$$

where $v_f = V_f/V_0$. From (3), (4), (5), and (9a), the relationship between $\theta_{\mathrm{off}}$ and $r$ is derived as

$$\frac{\pi r}{1 + v_f} = \tan \theta_{\mathrm{off}} - \theta_{\mathrm{off}}. \tag{11}$$

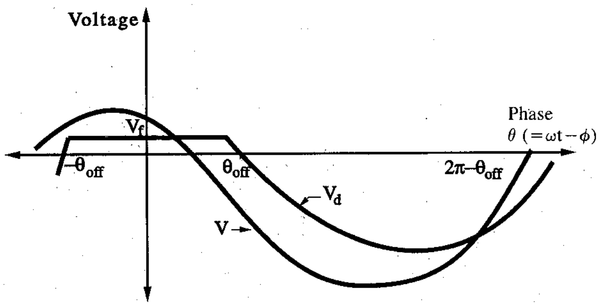


Fig. 4. Simplified time-domain waveform of voltage V and $V_d$. $V_d$ is the voltage between intrinsic diode. V is the voltage between A and C of Fig. 3. Variable $\theta$ is defined as $\theta = \omega t - \phi$.

$\theta_{\mathrm{off}}$ is determined by the ratio of $R_S$ to $R_L$ and the ratio of the forward voltage drop $V_f$ to the dc output voltage $V_0$. $V_f$ is solved with the $I$–$V$ relation of the diode:

$$I_s\left(\exp\left(\frac{eV_f}{nkT}\right)-1\right)=\frac{-V_0+V_1-V_f}{R_S} \tag{12}$$

$V_1$ in (12) is a function of $\theta_{\mathrm{off}}$ and therefore $V_f$ cannot be solved without first knowing $\theta_{\mathrm{off}}$. However, since the left term is of exponential form $V_f$ does not vary significantly from the cut-in voltage of the diode which is about 0.6 to 0.8 V for a GaAs-Pt Schottky diode. Therefore, $\theta_{\mathrm{off}}$ is solved from (11) substituting $V_f$ with a typical value of the cut-in voltage of the diode.

Once $\theta_{\mathrm{off}}$ is solved from a given $V_0$, $R_S$, and $R_L$, the efficiency and the input impedance can be calculated from the time domain waveform $V$ and $V_d$. These waveforms are expressed as a function of $\theta_{\mathrm{off}}$, diode parameters, and $V_0$ or $V_1$.

$$\begin{aligned}
P_{\mathrm{DC}} &= \frac{V_0^2}{R_L}\\
P_{\mathrm{loss}} &= \mathrm{LOSS}_{\mathrm{on},R_S}+\mathrm{LOSS}_{\mathrm{on,diode}}\\
&\quad+\mathrm{LOSS}_{\mathrm{off},R_S}+\mathrm{LOSS}_{\mathrm{off,diode}}\\
\mathrm{LOSS}_{\mathrm{on},R_S} &= \frac{1}{2\pi}\int_{-\theta_{\mathrm{off}}}^{\theta_{\mathrm{off}}}\frac{(V-V_f)^2}{R_S}\,d\theta\\
\mathrm{LOSS}_{\mathrm{on,diode}} &= \frac{1}{2\pi}\int_{-\theta_{\mathrm{off}}}^{\theta_{\mathrm{off}}}\frac{(V-V_f)V_f}{R_S}\,d\theta\\
\mathrm{LOSS}_{\mathrm{off},R_S} &= \frac{1}{2\pi}\int_{\theta_{\mathrm{off}}}^{2\pi-\theta_{\mathrm{off}}}\frac{(V-V_d)^2}{R_S}\,d\theta\\
\mathrm{LOSS}_{\mathrm{off,diode}} &= \frac{1}{2\pi}\int_{-\theta_{\mathrm{off}}}^{2\pi-\theta_{\mathrm{off}}}\frac{(V-V_d)V_d}{R_s}\,d\theta\\
\mathit{efficiency} &= \frac{P_{\mathrm{DC}}}{P_{\mathrm{loss}}+P_{\mathrm{DC}}}
\end{aligned} \tag{13}$$

where the variable $\theta$ is substituted for $\omega t - \phi$. The current ($I$) flowing through $R_S$ can be expressed as follows:

$$\begin{aligned}
I &= I_0 + I_{1r}\cos(\omega t) + I_{1i}\sin(\omega t)\\
I_0 &= \frac{1}{2\pi R_S}\left\{\int_{-\theta_{\mathrm{off}}}^{\theta_{\mathrm{off}}}(V-V_f)\,d\theta + \int_{\theta_{\mathrm{off}}}^{2\pi-\theta_{\mathrm{off}}}(V-V_d)\,d\theta\right\}\\
I_{1r} &= \frac{1}{\pi R_S}\left\{\int_{-\theta_{\mathrm{off}}}^{\theta_{\mathrm{off}}}(V-V_f)\cos(\theta+\phi)\,d\theta + \int_{\theta_{\mathrm{off}}}^{2\pi-\theta_{\mathrm{off}}}(V-V_d)\cos(\theta+\phi)\,d\theta\right\}\\
I_{1i} &= \frac{1}{\pi R_S}\left\{\int_{-\theta_{\mathrm{off}}}^{\theta_{\mathrm{off}}}(V-V_f)\sin(\theta+\phi)\,d\theta + \int_{\theta_{\mathrm{off}}}^{2\pi-\theta_{\mathrm{off}}}(V-V_d)\sin(\theta+\phi)\,d\theta\right\}.
\end{aligned} \tag{14}$$

The input admittance of the diode at the fundamental frequency is calculated as

$$Y=\frac{I_{1r}-jI_{1i}}{V_1}. \tag{15}$$

If the susceptance part resonates with the dc pass filter in Fig. 1, the input impedance to be matched to the antenna becomes the inverse of real part of $Y$.

To validate the closed-form equation the conversion efficiency calculated with the closed-form equation was compared with the results simulated by a nonlinear circuit computer program (LIBRA). The efficiency of the diode as defined by the ratio of output dc power to the net power transmitted to the rectenna is compared in Fig. 5 for 10 and 35 GHz. The diode parameters used for the calculation are summarized in Table I. Since the effect of $V_{br}$ was not included in the closed-form equation, the efficiency calculation using the closed-form equation terminated when the negative peak of the diode junction voltage reached $V_{br}$. At 10 GHz the efficiency calculated with the closed-form equation is close to the efficiency calculated with LIBRA. At 35 GHz there is relativly big difference in efficiency showing approximatly 10% of difference in the region where the efficiency calculated with LIBRA is not affected by the finite $V_{br}$. Increase of the difference at 35 GHz is considered to be caused by the effect of nonlinear junction capacitance which becomes dominant at the high frequency. The closed-form equation is not accurate enough to be used for the design of the high frequency rectenna. However, it can be used to estimate the preliminary efficiency of rectenna with any kind of diode at any arbitrary frequency. This is important because it predicts the efficiency level achievable through the nonlinear circuit optimization which usually takes a long time. It also explains how the efficiency is decreased by breaking down the power dissipation mechanism in the diode. In the next subsection we calculated the power conversion efficiency of an ideal diode using the closed-form equation.

### C. Power Conversion Efficiency of Ideal Diode

The power conversion efficiency of an ideal diode with a finite $R_S$ and $C_j$ was calculated using the closed-form

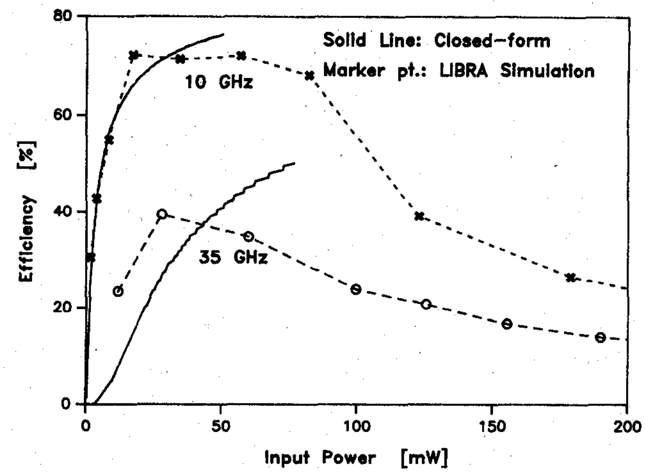


Fig. 5. Comparison of efficiencies calculated with closed-form formula and LIBRA. The diode parameters used for the calculation are listed in Table I. The dc load resistance is 400 Ω.

TABLE I
DIODE PARAMETERS OF THE *KA*-BAND GaAs BEAM-LEAD SCHOTTKY DIODE

| $V_{br}$ (volts) | $R_S$ (ohms) | $C_{jo}$ (pF) | $I_S$ (pAmps) | $\gamma$ | $C_p$ (pF) | $L_p$ (nH) |
|---|---|---|---|---|---|---|
| 9.0 | 4.85 | 0.13 | 5.24 | 0.5 | 0.1 | 0.2 |

$V_{br}$, breakdown voltage; $R_S$, parasitic series resistance; $C_{jo}$, zero bias junction capacitance; $I_S$, saturation current; $\gamma$, exponent in junction capacitance equation

$$C_j = \frac{C_{jo}}{\left(1 - \frac{V}{V_{bi}}\right)^{\gamma}}$$

$C_p$, $L_p$, parasitic reactive elements.; $V_{bi}$ is the built-in potential.

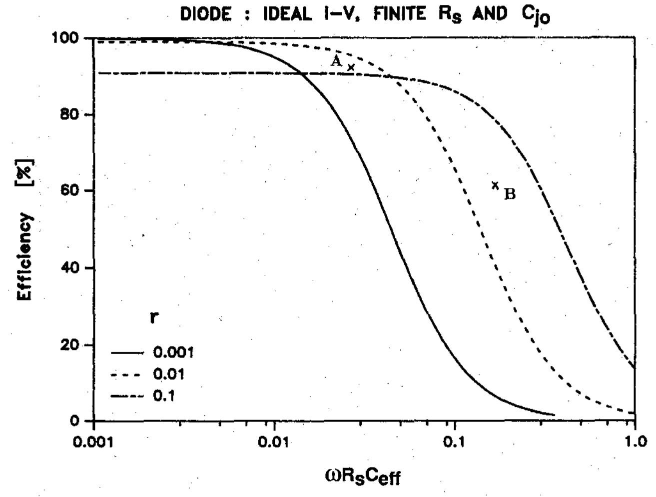


Fig. 6. Microwave-to-dc power conversion efficiency of an ideal diode with finite $R_S$ and $C_j$.

equation. The ideal diode here means the diode that an infinite current flows when the diode voltage is positive. The efficiency in Fig. 6 was calculated as a function of $C_{\text{eff}}$ for several different values of $r(= R_S/R_L)$. The x-axis variable is normalized by multiplying $C_{\text{eff}}$ by $\omega R_S$.

Since the efficiency of an ideal diode is the highest efficiency that can be achievable with an actual diode having the same $R_S$ and $C_j$ and an infinite $V_{br}$, Fig. 6 can be used as a preliminary estimation of the maximum efficiency of a diode at any arbitrary frequency. Point A in Fig. 6 corresponds to the efficiency of a diode ($R_S$ = 0.5 Ω, $C_{jo}$ = 3 pf) used for a 2.45 GHz rectenna with 100 Ω dc load and point B shows the efficiency of a *Ka*-band mixer diode ($R_S$ = 4.85 Ω, $C_{jo}$ = 0.13 pF) with 100 Ω dc load. Therefore the maximum efficiency of a 35 GHz rectenna is approximately 60%. The actual maximum efficiency will be smaller than 60% due to a finite $V_{br}$ and $V_f$.

## III. DESIGN AND MEASUREMENT OF 10 GHz AND 35 GHz RECTENNAS

### *A. Design Procedure*

Due to the nonlinearity of the rectenna the operating power greatly affects the circuit performance. Therefore the design of the rectenna starts by determining the operating power level. The design procedure is thus as follows:

1) The dc output power ($P_{\text{out}}$) is decided and the dc voltage $V_0$ is designed to be less than half of $V_{br}$.
2) The dc load resistance is calculated from the equation of $R_L = V_0^2/P_{\text{out}}$ with $P_{\text{out}}$ and $V_0$ determined in step (1).
3) The effective admittance of the diode is calculated from the closed-form equation or a nonlinear circuit simulation with $V_0$ and $R_L$ values determined in steps (1) and (2).
4) Design the antenna and calculate the input impedance.
5) The low pass filter is designed with the input impedance of the antenna and the effective impedance of diode. If the dc pass filter is designed to resonate with the imaginary part of the diode admittance, the inverse of the real part becomes the effective input impedance of the diode. This impedance is matched to the antenna through the low pass filter.
6) The efficiency is maximized through the nonlinear computer simulation.

### *B. 35 GHz Dipole Rectenna*

A GaAs Schottky diode (Model DMK6606, Alpha Industries) used in *Ka*-band mixers was selected for the 35 GHz rectenna. The diode parameters are summarized in Table I. A 35 GHz rectenna was designed using a microstrip dipole as the antenna element. The input impedance of the dipole as an element in an infinite array was calculated using Method of Moment [14]. The impedances calculated from the fundamental frequency to the seventh order harmonic were used as an input data file to LIBRA. The length and the width of a resonant dipole at the fundamental frequency have been determined as 0.46 $\lambda_o$ and 0.02 $\lambda_o$ respectively. The effective impedance of the diode was assumed to be 50 Ω considering that the output dc voltage should be less than 4.5 V (half of $V_{br}$) with an operating power level of 100 mW.

The low pass filter which consists of three transmission line sections was optimized using TOUCHSTONE. The cutoff frequency of the low pass filter was located between the fundamental and the second order harmonic. Because the coplanar strip line was not available in TOUCHSTONE, an ideal transmission line was substituted for the optimization. The ideal transmission line was converted to the physical transmission line using the closed-form equation in [15]. The dimensions of the transmission lines are shown in Table II. The Duroid substrate was 10 mils thick with a 2.2 dielectric constant.

The efficiency and the reflected power have been calculated through the nonlinear circuit simulation (LIBRA). The circuit model for the simulation is shown in Fig. 7. The efficiency was maximized by adjusting *L4* of the dc pass filter. The optimum length of *L4* was 0.022 $\lambda_o$. The circuit configuration and the photograph of the actual circuit are shown in Fig. 8. The circuit was fabricated on a substrate having a thin (10 $\mu$m) copper layer because the smallest gap of the circuit was 100 $\mu$m. This thickness is still 30 times larger than the substrate skin depth at 35 GHz. The circuit was etched using the conventional method and a beamlead diode was mounted on the circuit with a silver epoxy.

### *C. 35 GHz Rectenna Measurements*

The overall efficiency is defined as the ratio of the output dc power to the incident power. The incident power includes both input power to the rectenna diode and reflection power. A waveguide array simulation technique [16] was used to measure the overall efficiency. This technique permits the generation of high power densities and the accurate measurement of efficiency. The measurement setup is shown in Fig. 9. A special waveguide expander was fabricated to appropriately simulate the rectenna array. The smaller side of a *Ka*-band waveguide (WR-28) was enlarged to make the cross section at the end of the waveguide expander be square. The longer side was not expanded in order for the mainlobe of the simulated array to be located in the same direction as that of two plane waves constructing the $TE_{10}$ mode. The area of the expanded cross section was 0.28 inch by 0.28 inch. This area is 40% larger than the effective area of a dipole antenna located a quarter wavelength above a ground plane. Therefore, the actual reflection of the rectenna might be smaller than the value measured with this setup. A photo of waveguide expanded section, a quarter wavelength spacer and a reflecting ground plane are shown in Fig. 10. The quarter wavelength spacer was placed between the rectenna and the reflecting plane.

The measured data is plotted in Fig. 11. The circles show the measured efficiencies which are defined as the ratio of the dc output power to the input power into the diode. The simulated results have been plotted together for comparison. The theoretical efficiency with 100 Ω dc loads are better than those with 400 Ω loads. The highest

TABLE II
OPTIMIZED LENGTHS OF COPLANAR STRIP LINES OF 35 GHz RECTENNA

| L1 (mm) | L2 (mm) | L3 (mm) | L4 (mm) |
|---|---|---|---|
| 0.92 | 1.19 | 0.58 | 0.18 |

$Z_{high}$: 245 Ω Width: 0.2 mm Gap: 0.4 mm
$Z_{low}$: 130 Ω Width: 0.5 mm Gap: 0.1 mm

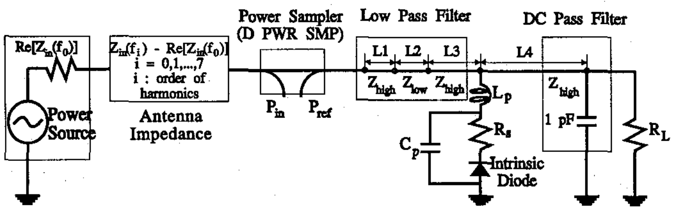
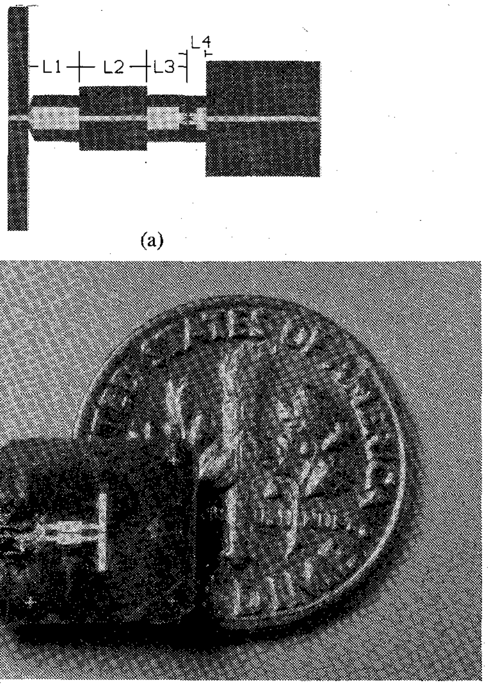


Fig. 7. Circuit model of 35 GHz dipole rectenna for the nonlinear circuit computer simulation (LIBRA). $Z_{in}(f_i)$: Antenna input impedance at the *i*th harmonic frequency. The real part of the antenna input impedance is used as a source resistance for the power source. The dual port power sampler (DPWRSMP) is used to measure the input and the reflected power.



(a)

(b)

Fig. 8. 35 GHz dipole retenna. (a) Circuit configuration. (b) Photograph of an actual circuit.

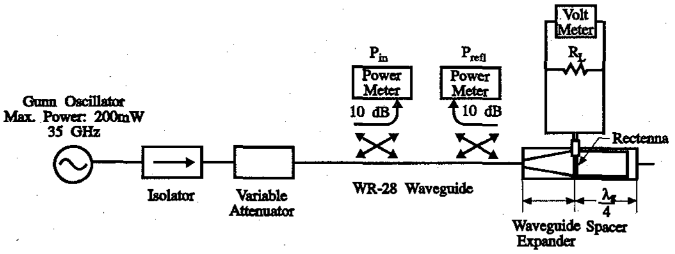


Fig. 9. Experimental setup for the measurement of the overall efficiency and the reflected power of a 35 GHz rectenna.

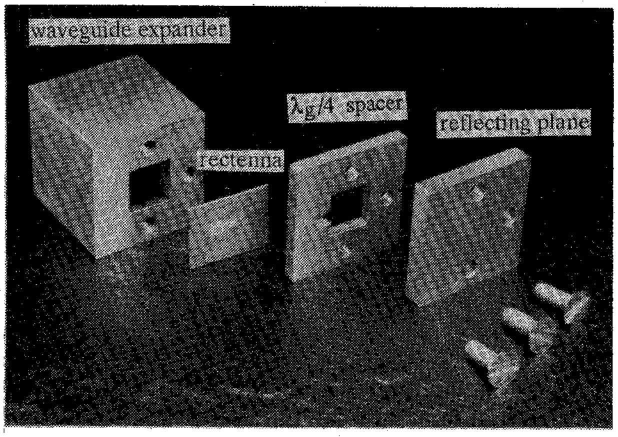
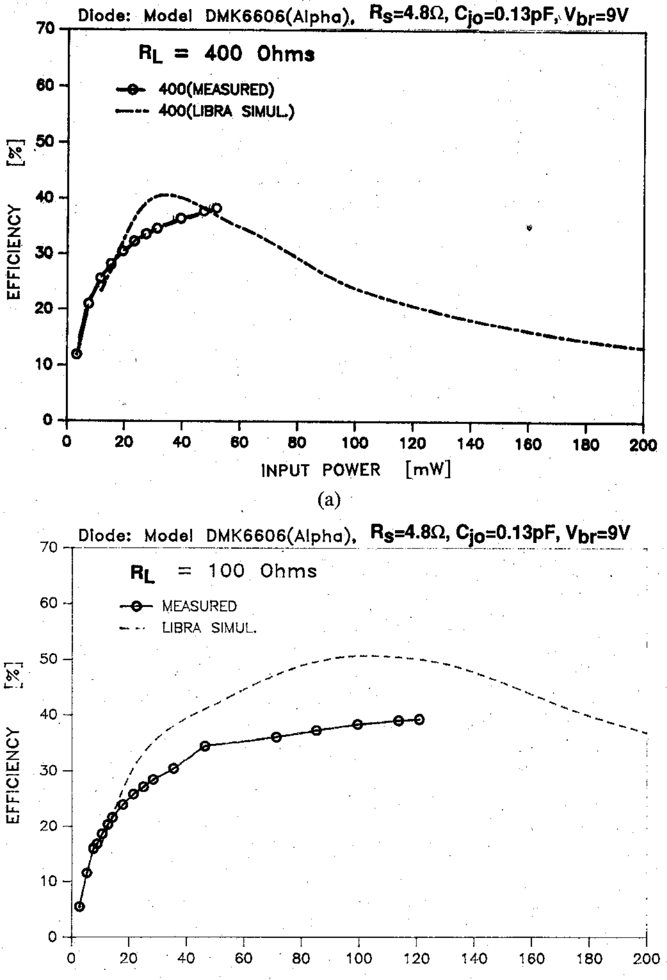


Fig. 10. A waveguide expanded section assembly consisting of a waveguide expander, a quarter wavelength spacer, and a reflecting plane.



(a)

(b)

Fig. 11. 35 GHz to dc power conversion efficiency. (a) With a 400 Ω dc load. (b) With a 100 Ω dc load. The efficiencies calculated by LIBRA are plotted together to compare with the measured result. The diode parameters used for simulation are listed in Table I.

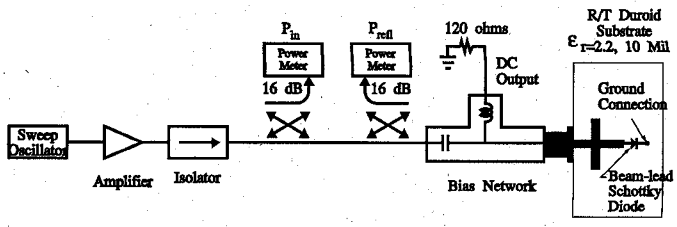


Fig. 12. Experimental setup for the measurement of the conversion efficiency of a *Ka*-band beam-lead diode at 10 GHz.

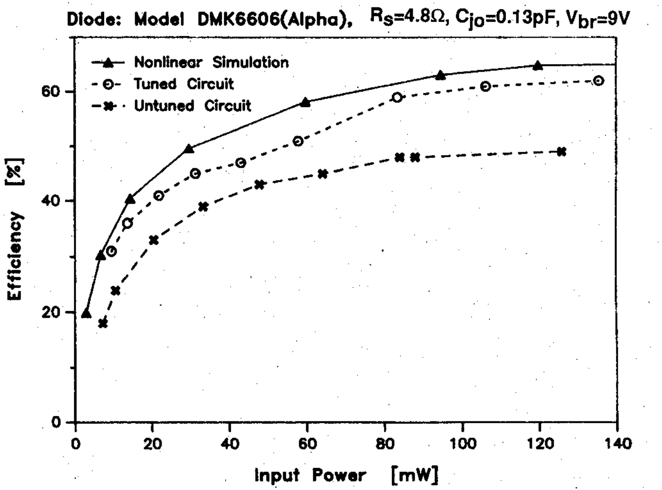


Fig. 13. 10 GHz to dc conversion efficiency of a *Ka*-band diode measured with 120 Ω dc load.

efficiency measured was 39%. The measured maximum efficiencies for 100 and 400 Ω loads are about the same. But the reflection power is 12% of the incident power at 400 Ω load and 21% of the incident power at 100 Ω load. The reflection can be reduced in the array environment.

### D. Measurement of Power Conversion Efficiency at 10 GHz

The power conversion efficiency of a *Ka*-band diode (Model DMK6606, Alpha Industries) was measured at 10 GHz. A coaxial cable directional coupler and 50 Ω coaxial cables were used in the measurement as shown in Fig. 12. The circuit consisted of a 50 Ω microstrip transmission line and a beam-lead diode. One end of the beam-lead diode is connected to the end of microstrip line and the other end is connected to the ground plane through a hole. The dc output power was extracted using a bias-T network. The efficiency was measured with and without the open stub tuner illustrated in Fig. 12. The measured efficiencies are plotted in Fig. 13. There was approximatly 10% of improvement with the open stub tuner. The efficiency with the open stub tuner approaches 60% which is agreed with the results from nonlinear computer simulation results.

## IV. CONCLUSION

A 35 GHz rectenna has been developed with 39% of power conversion efficiency. The rectenna used a microstrip dipole and a *Ka*-band mixer diode. A computer program LIBRA was used to optimize the rectenna for the high conversion efficiency. The net power conversion efficiency of the same diode was measured at 10 GHz using a simple rectifying circuit. The net conversion efficiency was 60%. If the impedance of a antenna is matched to a rectifying circuit, 60% efficiency would be possible at GHz with this diode. A closed-form equation was derived for the analysis of the high frequency rectenna by neglecting the higher order harmonics. In order to estimate the maximum efficiency limited by a finite $R_S$ and $C_{jo}$, a net conversion efficiency of an ideal diode with the same $R_S$ and $C_{jo}$ was calculated using the closed-form equation. With the *Ka*-band diode used for 35 GHz rectenna, the maximum efficiency was calculated to be approximatly 60% at 35 GHz. The actual maximum efficiency is smaller than this efficiency due to the effect of the finite forward voltage drop and the breakdown voltage.

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## REFERENCES

[1] W. C. Brown *et al.*, U.S. Patent 3 434 678, Mar. 25, 1969.

[2] W. C. Brown, "Experiment in the transport of energy by microwave beam," in *IEEE Int. Conv. Rec.*, vol. 12, pt. 2, Mar. 1964, pp. 8-18.

[3] —, "The history of power transmission by radio waves," *IEEE Trans. Microwave Theory Tech.*, vol. 32, no. 9, pp. 1230-1242, Sept. 1984.

[4] —, "Experiments involving a microwave beam to power and position a helicopter," *IEEE Trans. Aerosp. Electron. Syst.*, vol. AES-5, no. 5, pp. 692-702, Sept. 1969.

[5] —, "Solar Power Satellite Program Rev. DOE/NASA Satellite Power System Concept Develop. Evaluation Program," Final Proc. Conf. 800491, July 1980.

[6] J. Schlesak, A. Alden and T. Ohno, "A microwave powered high altitude platform," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1988, pp. 283-286.

[7] W. C. Brown, "Performance characteristics of the thin-film, etched-circuit rectenna," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1984, pp. 365-367.

[8] J. J. Nahas, "Modeling and computer simulation of a microwave-to-dc energy conversion element," *IEEE Trans. Microwave Theory Tech.*, vol. 23, no. 12, pp. 1030-1035, Dec. 1975.

[9] W. C. Brown and E. E. Eves, "A microwave beam power transfer and guidance system for use in an orbital astronomy support facility," Final Report (Phase II) to NASA, NAS-8-25374, Sept. 1972.

[10] G. P. Boyakhchyan *et al.*, "Analytical calculation of a high-efficiency microwave rectifier employing a Schottky-barrier diode," *Telecom. Radio Engin.* (English Translation of Elaktrosvyaz, and Radiotekhnika) vol. 37-38, no. 10, pp. 64-66, Oct. 1983.

[11] W. C. Brown, "All electronic propulsion key to future space ship design," presented at *AIAA/ASME/SAE/ASEE 24th Joint Propulsion Conf.*, Boston, July 1988.

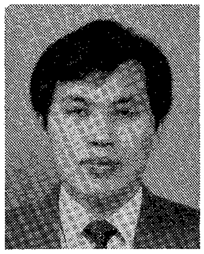
[12] K. Chang, J. C. McCleary, and M. A. Pollock, "Feasibility study of 35 GHz microwave power transmission in space," *Space Power*, vol. 8, no. 3, pp. 365-370, June 1989.

[13] S. M. Sze, *Physics of Semiconductor Devices*. New York: Wiley, 1981, ch. 2 and 5.

[14] S. M. Wright, "Efficient analysis of infinite microstrip arrays on electrically thick substrate," Ph.D. dissertation, University of Illinois at Urbana-Champaign, 1984.

[15] I. J. Bahl, "Transmission lines," in *Handbook of Microwave and Optical Components*, K. Chang Ed., vol. 1, New York: Wiley, 1989, p. 35.

[16] P. W. Hannan and M. A. Balfour, "Simulation of a phase-array antenna in a waveguide," *IEEE Trans. Antennas Propagat.*, vol. AP-13, pp. 342-353, May 1965.



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