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" Quasi-optical Transponder Using FETs and Patch Antennas "

By : Kimin Cha and Tatsuo Itoh
Electrical Engineering Department

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ABSTRACT

Quasi-optical planar transponder that integrates a self-oscillating mixer, an IF amplifier and two patch antennas, one for receiving RF signal (14 GHz) and another for transmitting amplified IF signal (8 GHz), is presented. The prototype circuit is designed in microstrip structure and fabricated on a single substrate. As an active source, GaAs FET is selected. Self-oscillating (6 GHz) mixer with patch antenna in receiver front end is developed based on quasi-optical technique. To obtain enough power level of IF signal before it is transmitted, the amplifier is designed with bandpass filter which blocks undesirable mixing products except 8 GHz. This prototype of transponder, which is compact and self-contained unit as a single entity has many applications in microwave and millimeter-wave area.

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Chapter 1

Introduction

Recently traffic congestion problems on freeways have increased. The problem is worse if there are toll-gates. Toll collector has traditionally required manual handling for each car. To solve this problem, a transponder mounted on the car which is read by interrogating transceivers positioned at the toll-gate will be useful. In many microwave and millimeter-wave applications, such as the transponder, it is essential to use planar circuit technology to reduce the size and cost of the systems. When one uses conventional discrete components to build a transponder, the system becomes impractical due to its large size, complexity, and cost. Over the past several years, the microwave monolithic integrated circuit (MMIC) technology has made tremendous progress which makes it possible to create a new form of active antennas, a part of our transponder, which consist of passive planar antenna elements and active semiconductor devices at higher frequencies. More compact, low cost, and light-weight transponders are essential for space and personal communication. Especially a small and conformal transponder is expected to be realized for mobile communication systems and contactless ID systems. Due to these requirements, the developments are necessary. The active antenna and quasi-optical technologies have been used to satisfy these requirements.

In this thesis, the transponder that integrates a self-oscillating mixer (SOM), an IF amplifier, and two patch antennas on the same substrate is developed. This circuit is compact and low power self-contained unit which is suitable for commercial applications. Due to its planar nature and compactness, the transponder can be fabricated by MMIC technology, and have potential to be a low cost unit.

This thesis consists of three main sub-sections (the receiving part, the transmitting part, and the transponder) which traces the steps on how the transponder is designed and fabricated. In Chapter 2, the receiving part of the entire circuit is described. Designing the SOM is the crucial step in the receiving part. The concept of the self-oscillating mixer was studied in Tajima's paper [1] and the demonstration of that using quasi-optical technique was presented in Hwang's paper [2]. Several researchers have demonstrated MMIC compatible planar quasi-optical mixers [3] - [5] which combine the functions of antenna and conventional mixer. These quasi-optical mixers reduce the size and complexity of the circuits considerably, however, they use an external local oscillator. In [3] the LO signal is fed via a waveguide. In [4] and [5] the LO signal is coupled through free space using a waveguide horn to radiate LO power to the mixers. Although this technique is simple, it involves substantial loss of the LO power. At millimeter-wave frequencies, one often cannot afford such a large power loss. Hence, the integration of a local oscillator and a mixer to form a self-contained unit should be needed. This concept has been extended to quasi-optical active transponder as described in this study. Combination of the quasi-optical technology and MMIC technology results in a variety of active circuits.

As the active device of this single entity, the FET is frequently selected for application of MMIC technology. Using the FET several active antennas and quasi-optical oscillators have already been reported which demonstrate topology useful for the MMIC technology [6]. Since the FET has high DC-RF conversion efficiency and is a three-terminal device, we can obtain the conversion gain in the mixer stage. The mixer stage of the transponder dominates the performance of the entire circuit, since the RF signal received by the antenna is fed directly to the mixer without going through an RF pre-amplifier. In Chapter 2, some fundamental concepts of the microwave mixers are introduced. The frequency mixing mechanism of the

MESFET mixers are described and their advantages and disadvantages are evaluated. In Chapter 3, the transmitting part which consists of a bandpass filter, an IF amplifier, and a patch antenna is described. The basic building block of this transponder is two patch antennas, one for receiving RF signal, and another for transmitting IF signal. Among the several advantages offered by microstrip patch antennas is the ease of their integration with microstrip circuits. In Chapter 3, how the patch antenna is chosen and designed is described. This transmitting part including the patch antenna, performing both as a load condition of the circuit and antenna functions in a single compact module, offer several distinct advantages including: (a) reduction in the substrate area and consequently, the size; (b) elimination of interconnect lines and connectors; (c) possibility for performance enhancement of combined circuit-antenna components; (d) power combining for oscillators, and (e) potential cost reduction.

As the last main part, in Chapter 4, the entire circuit of the transponder is presented. The most significant point in the design of this transponder is to incorporate the input impedance of the transmitting part and the receiving antenna into the model of the entire circuit. In the case of the active antenna using a microstrip line, the input impedance is denoted as a series impedance which depends on the frequency. In order to determine the configuration of the FET oscillator, a negative resistance value is set using small signal S-parameter. However, to evaluate the steady state oscillation frequency or other parameters under the entire circuit configuration including the patch antenna input impedance, a large signal analysis should be carried out. For this purpose, there are two ways; the harmonic balance analysis [7] and the time domain analysis. Both methods are available in the commercial-use microwave CAD. To obtain the information of the steady state oscillation, one of them is necessary.

Chapter 2

Receiving Part

2.1 Introduction

In this chapter, the receiving part of the quasi-optical transponder that employs a patch antenna, a MESFET self-oscillating mixer and a bandpass filter is described. Above all, the self-oscillating mixer (SOM) is the key component of this transponder. The FET SOM circuit offers three advantages. First, the new circuits exhibit conversion gain rather than conversion loss, as in the diode mixer circuits reported in [8]. Second, the new circuits are more compatible with the FET-based MMIC technology since no diodes are used. And third, fewer devices are required to construct a simple receiver front end. Using this SOM a compact and self-contained transponder can be designed which has potential applications in microwave and millimeter-wave area. However, the difficulty with self-oscillating mixers is to obtain sufficient oscillation amplitude at a bias condition where nonlinearity is the strongest.

A mixer performs the function of frequency conversion. This frequency conversions are carried out by signal multiplication as shown in Fig.2.1. In practice, an ideal signal multiplying device does not exist. The mixing function is performed by nonlinearity of devices. This frequency conversion can be explained by the I-V characteristic of a nonlinear device via a power series:

$$I=a_0+a_1V+a_2V^2+a_3V^3+\cdots \qquad (2.1)$$

where a_0, a_1, and a_2 are constant real coefficients. Let V be a two-tone excitation (the sum of the RF and LO voltage) of the form:

$$V=V_1\cos(\omega_1 t) + V_2\cos(\omega_2 t) \qquad (2.2)$$

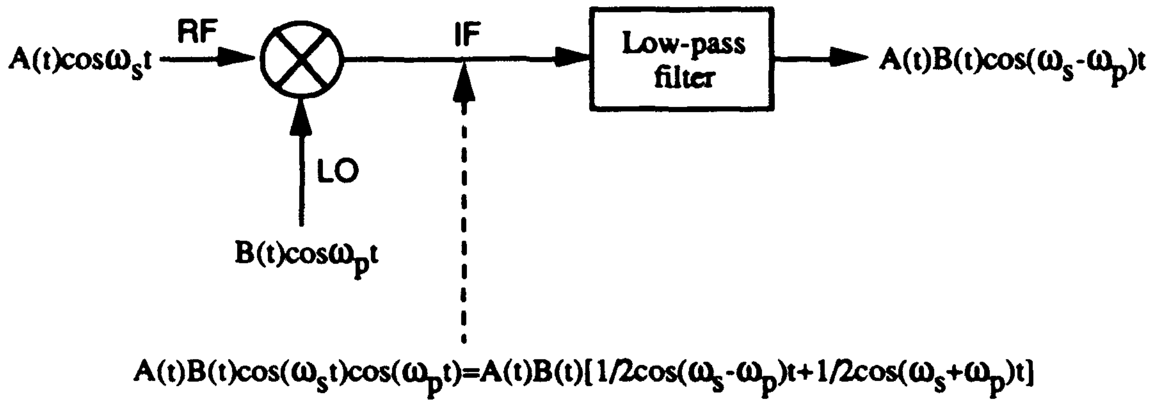

Fig. 2.1 Ideal signal Multiplier

Substituting (2.2) into (2.1) gives, for the first term:

$$i_0(t)=a_0V_1\cos(\omega_1 t) + a_0V_2\cos(\omega_2 t) \qquad (2.3)$$

After doing the same with the second term and applying the well-known trigonometric formulas for squares and products of cosines, we obtain

$$\begin{aligned} i_1(t)=\frac{1}{2}a_1\{&V_1^2+V_2^2+V_1^2\cos(2\omega_1 t)+V_2^2\cos(2\omega_2 t) \\ &+2V_1V_2[\cos((\omega_1+\omega_2)t)+\cos((\omega_1-\omega_2)t)]\} \end{aligned} \qquad (2.4)$$

and the third term gives

$$\begin{aligned} i_2(t)=\frac{1}{2}a_2\{&V_1^3\cos(3\omega_1 t)+V_2^3\cos(3\omega_2 t) \\ &+3V_1^2V_2[\cos((2\omega_1+\omega_2)t)+\cos((2\omega_1-\omega_2)t)] \\ &+3V_1V_2^2[\cos((2\omega_2+\omega_1)t)+\cos((2\omega_2-\omega_1)t)] \\ &+3(V_1^3+2V_1V_2^2)\cos(\omega_1 t) \\ &+3(V_2^3+2V_1^2V_2)\cos(\omega_2 t)\} \end{aligned} \qquad (2.5)$$

Therefore, we can expect the total current contains the terms at the frequencies of

$$\omega = m\omega_{RF} \pm n\omega_{LO} \qquad \text{for } m,n=0,1,2,3,... \qquad (2.6)$$

The frequency $\omega_{RF} \pm \omega_{LO}$ is called the first order mixing product. The rest of terms in (2.6) are the harmonics of the LO and RF signals and the intermodulation products.

2.2 Self-Oscillating Mixer

2.2.1 MESFET Mixer

The primary device for microwave mixers from below 1 GHz to above 300 GHz is the Schottky-barrier diode. Nevertheless, the successful device development of low-noise GaAs MESFETs, which operate into the millimeter-wave region, has created the possibility of designing FET mixers with performance equal or superior to that of diode mixers at frequencies as high as 50 GHz [9]. FET mixers are capable of achieving noise and intermodulation lower than diode mixers, with less LO power.

Work on MESFET mixers has been reported by many researchers. The pertormance of the single-gate MESFET mixer was first investigated by Pucel et. al. [10]. In Pucel's work, it was shown that the MESFET mixer can provide the conversion gain, high dynamic range, and low noise. Also, the MESFET mixer is particularly well suited for MMIC. There are three basic types of MESFET mixers, the gate mixer [11]-[15], the drain mixer [16], and the resistive mixer [17].

In the gate mixer, both the LO and RF signals are applied to the gate of a MESFET. The device is biased to near pinch-off (Fig.2.2) where the device's transconductance is most sensitive to LO modulation. The drain current, which is the product of the RF input voltage and the transconductance, includes the IF term if expanded by power series. This drain current is increased by pushing up the gate voltage which is modulated by the LO signal. Since the RF and LO signals are both applied to the gate in a single gate MESFET gate mixer, a diplexer is required to isolate the RF and LO ports. This is the major disadvantage of the single-gate MESFET mixer since the diplexer increases the complexity and size of the circuit.

In the drain mixer, the RF signal is applied to the gate, and the LO signal is applied to the drain. The device is drain-biased near the knee voltage and is gate-biased to a small negative voltage or zero (Fig.2.2). At this bias condition, the device output resistance R_o, and transconductance G_m, are both nonlinear, and the device voltage amplification factor $u=G_m*R_o$ is modulated by the LO signal. Thus, the voltage at the drain port, $V_o=V_{RF}*u$, will contain a component at the IF frequency.

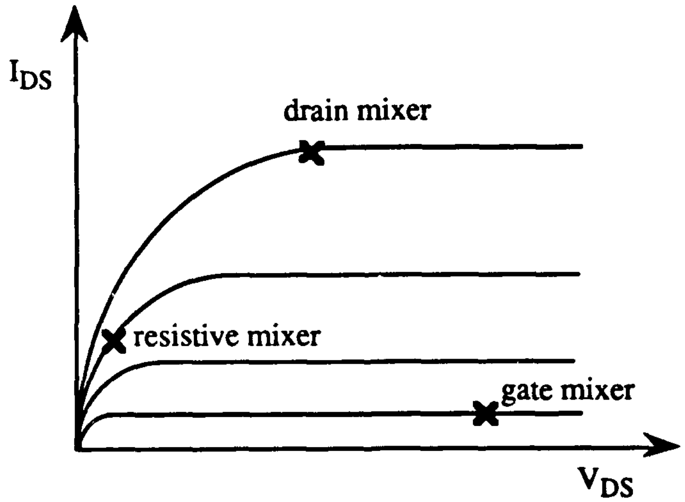

Fig. 2.2 Bias points of the gate and drain FET mixer

The MESFET can also be used as a resistive mixer. Maas [17] reported a GaAs MESFET mixer with very low intermodulation products. In the resistive mixer, the channel of the MESFET is used to realize a time-varying resistance which is modulated by the LO voltage applied to the gate. Because of the weak nonlinearity of the gate voltage to channel resistance relationship, the mixer generates very low intermodulation products and is capable of high output power at moderate LO levels. The MESFET resistive mixer operates in the unsaturated (linear) region of the device's I-V characteristic as shown in Fig.2.2. The LO voltage is applied to the gate of a MESFET where the unbiased channel operates as a simple resistor whose conductance, $G(V_{LO})$, is modulated by the LO signal while the RF signal is applied to the drain port. The current at the drain port, which is equal to $I_D=V_{RF}*G(V_{LO})$ contains the IF frequency term if it is expanded into a power series.

In the present work, a MESFET SOM is incorporated with the patch antenna to form a self-contained receiver. Since the SOM is adopted, the mixer is modified to be able to operate as an oscillator simultaneously. In this SOM, the device is drain-biased and gate-biased near pinch-off. This bias condition is similar to that of the gate mixer. The drain current, which is the product of the RF voltage and the transconductance, includes the IF signals.

2.2.2 Self-Oscillating Mixer Design

Interest in microwave and millimeter-wave SOM has been on the increase in recent years, mainly because of the power efficiency, low cost, and comparatively simple circuitry for communicational applications. In conventional way, we need two components to realize the mixing function: a local oscillator (LO) and a mixer. It should be noted that the SOM does not need a separate LO and mixer. It acts simultaneously as a local oscillator and a mixing element. The first step in the design of a SOM is the design of the circuit as an oscillator. We need to study how the oscillator is designed first.

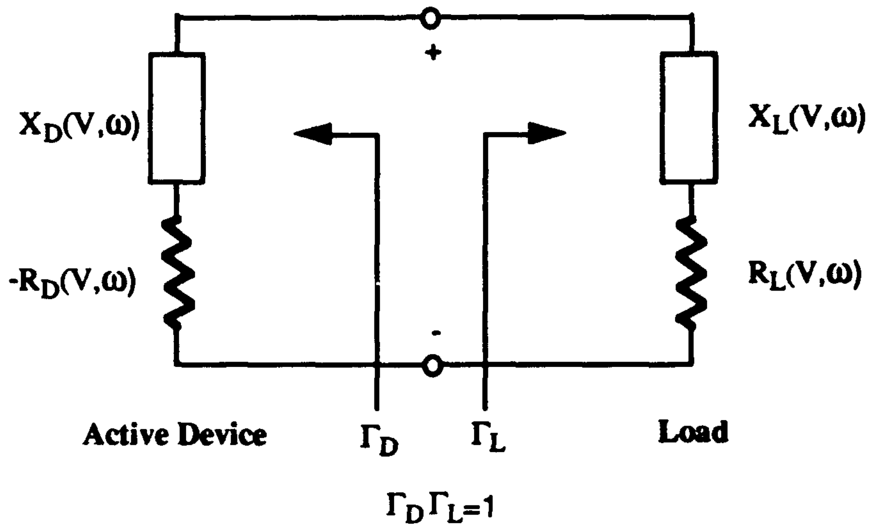

Fig. 2.3 1-port Negative Resistance Oscillator

There are two types of solid state oscillators. One is a feedback type and the other a negative resistance type. In the former, when the loop gain of the feedback equals one, the net gain of the feedback oscillator becomes infinite. Hence, the circuit operates in steady state oscillation. The negative resistance approach [18], however, is generally applied to a one-port negative resistance oscillator as shown in Fig.2.3. The negative resistance device is represented by the amplitude and frequency-dependent impedance.

For the steady state oscillation, the voltage around the loop is zero. Using the parameters shown in Fig.2.3,

$$R_D(V,\omega) + jX_D(V,\omega) + R_L + jX_L(\omega) = 0 \qquad (2.7)$$

The oscillation conditions can be expressed in the form

$$\begin{aligned} -R_D(V,\omega) + R_L &= 0 \\ X_D(V,\omega) + X_L(\omega) &= 0 \end{aligned} \qquad (2.8)$$

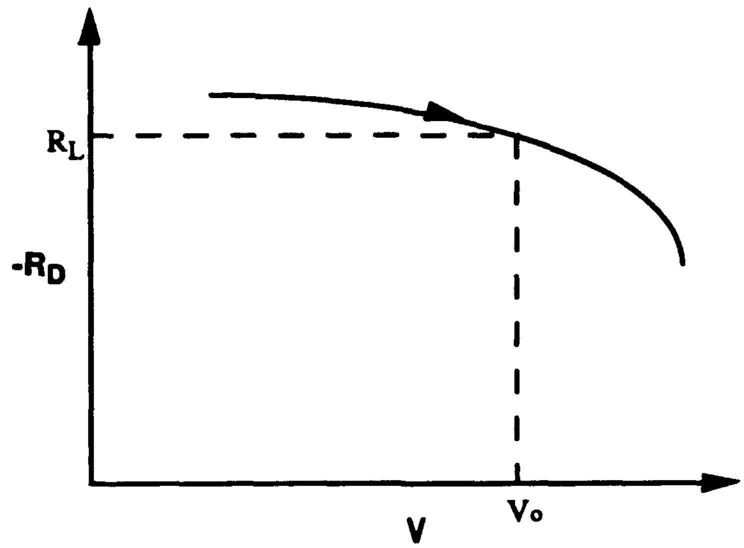

Fig. 2.4 Negative Resistance Variation of Active Device in Series Circuit Model

In (2.8), R_D and X_D have the amplitude and frequency dependence. By using the reflection coefficient, the steady state oscillation condition is also expressed as $\Gamma_D\Gamma_L=1$. When the amplitude V increases $R_D(V,\omega)$ decreases as shown in Fig.2.4. Therefore, in order for the circuit to oscillate, R_L is required to be less than the initial $R_D(V_o,\omega_o)$. This means that the one-port network is unstable if the net resistance of the network is negative $(-R_D(V,\omega) + R_L < 0)$.

The frequency of oscillation determined by (2.8) is not stable since $X_D(V,\omega_o)$ is amplitude dependent. It is necessary to find another condition to guarantee a stable oscillation. If the frequency dependence of $X_D(V,\omega)$ can be neglected for small variation around ω_o, Kurokawa [19] has shown that the condition for a stable condition is

$$\left.\frac{\partial R_D(V,\omega)}{\partial V}\right|_{V=V_o}\left.\frac{\partial X_L(\omega)}{\partial \omega}\right|_{\omega=\omega_o} - \left.\frac{\partial X_D(V,\omega)}{\partial V}\right|_{V=V_o}\left.\frac{\partial R_L(\omega)}{\partial \omega}\right|_{\omega=\omega_o} > 0 \qquad (2.9)$$

That is, the frequency of oscillation determined by (2.8) is stable only if (2.9) is satisfied. Although one can obtain the steady state oscillation through the technique described above, this may not be sufficient for a stable operation condition. To evaluate the steady state oscillation frequency, a large signal analysis need to be carried out. For this purpose, there are two methods. One is the harmonic balance analysis and the other the time domain analysis. Both methods are available in the commercial-use microwave CAD (mwSPICE, Libra,...)

2.3 Measurement Results

The receiving part including the SOM and the bandpass filter was designed and fabricated as shown in Fig.2.5. ROGERS RT/Duroid 5870 (ε_r=2.33, thickness=31 mil, 1 oz Cu sheet) as the substrate and the package-type FET (Avantek ATF26884) as the active source were used. Since mixing is automatically happened by the nonlinear characteristics of the device, to design the negative resistance oscillator should be more crucial.

In the receiving part, the higher order mixing product which only could pass through the bandpass filter was concentrated on. With 10 GHz RF signal and 6 GHz LO signal one could expect many mixing signals, especially 4 GHz and 16 GHz which are the first order mixing products. Among them 8 GHz was picked up which is the second harmonic of 4 GHz or the mixing product from RF and the third harmonic of the oscillation frequency. There were two reasons. One was how much the power of the higher order mixing products could be obtained with this SOM. Another was that only one X-band standard gain horn in experimental setup was preferred to use for simple setup. It was successful as shown in Fig.2.7. However, the power level of that signal was too small to be detected (at most 1 ft distance).

Hence, the transponder was redesigned to pick up the first order mixing signal, which is described in Chapter 4. The patch antenna was designed at 10 GHz (RF signal) of which the input impedance was 248.5 Ω at the feed point. Including the input impedance of the patch antenna many oscillator circuits were designed at 6 GHz (LO signal) and tested for determining the schematic of the SOM as well as considering the oscillation power. The measured oscillation power directly in the spectrum analyzer without going through the bandpass filter was 5 dBm at 6 GHz.

Since 8 GHz is the higher order mixing product, the signal is expected to be very low. To obtain enough signal level to detect this signal LO signal (6 GHz) and other signals except 8 GHz should be kept inside the circuit as much as possible and designed the optimum impedance matching for 8 GHz. Therefore, the bandpass filter which is described in section 3.1 was utilized to block any other signals except 8 GHz.

Including the impedance of the patch and the bandpass filter the SOM was designed at 6 GHz. At the reference plane in Fig.2.5 the load impedance of this negative resistance oscillator was 1.296 + j 91.031 Ω of which the real part was fairly small. Using the load impedance of the SOM R_D was set approximately three times of the real part of the load impedance based on the theory in [18]. The simulation result shows the oscillation frequency in Fig.2.6. Normally there is 10% difference between the oscillation frequency of the simulation and that of the experiment. Hence, the designed frequency in the simulation was set around 6.7 GHz as seen in Fig. 2.6. For a design tool the EEsof Touchstone® was used.

The measurement results are in Fig.2.7 and Fig.2.8. RF was at 9.96 GHz which was generated by the sweep oscillator and LO signal was produced at 5.88 GHz. The second harmonic of the first order mixing product was measured exactly at 8.16 GHz (= 2•(9.96-5.88)). As expected this signal level was very low. From the RF and the third harmonic of the oscillation frequency, a strong signal (7.76 GHz = 17.72-9.96) compared with 8.16 GHz was detected. After tuning the RF signal two signals were synchronized as shown in Fig.2.8. The receiving antenna pattern was measured as in Fig.2.9.

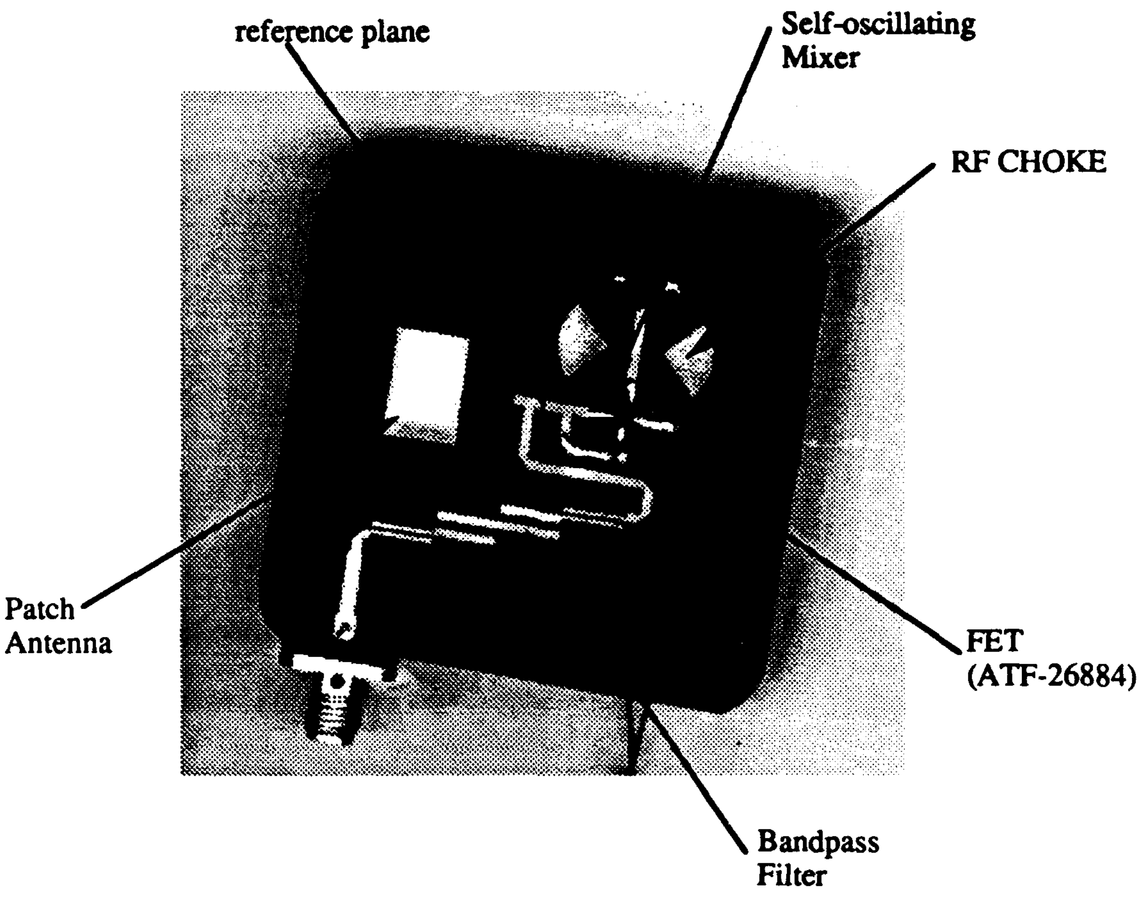

Fig. 2.5 Receiving Part of Transponder

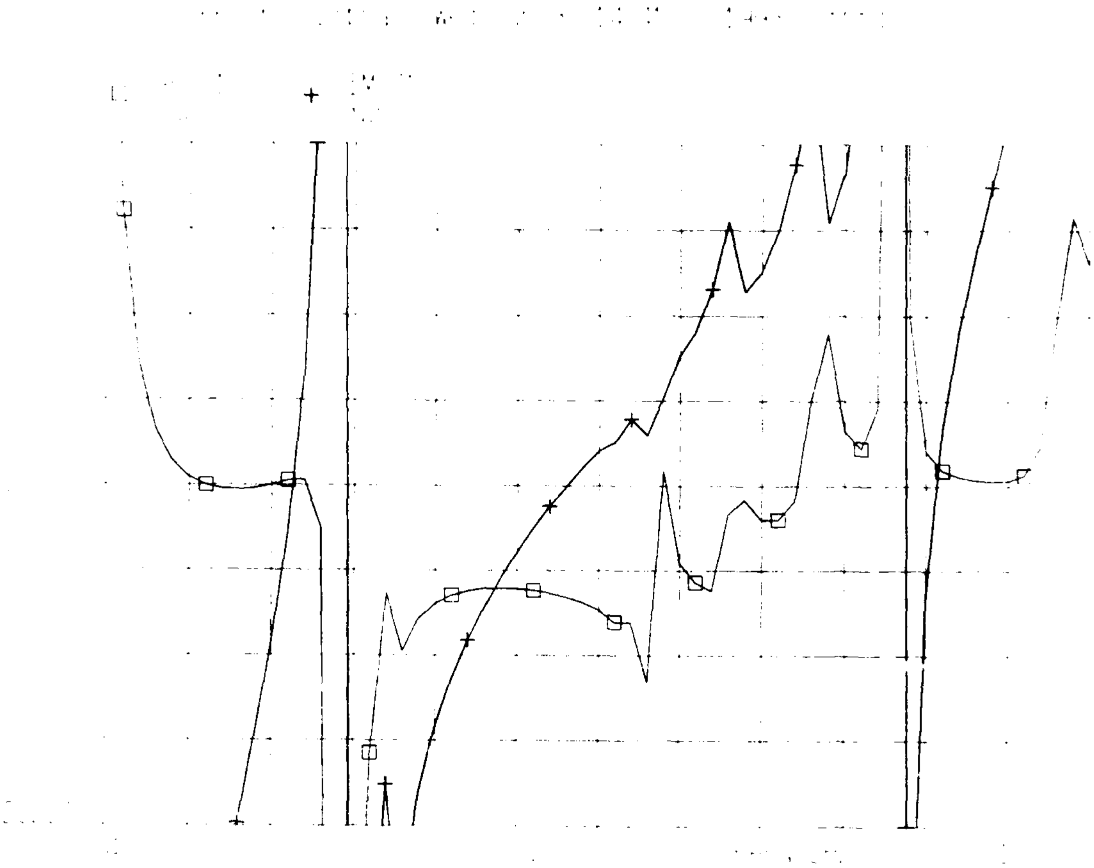

Fig. 2.6 Simulation Result of the negative resistance Oscillator using small signal analysis

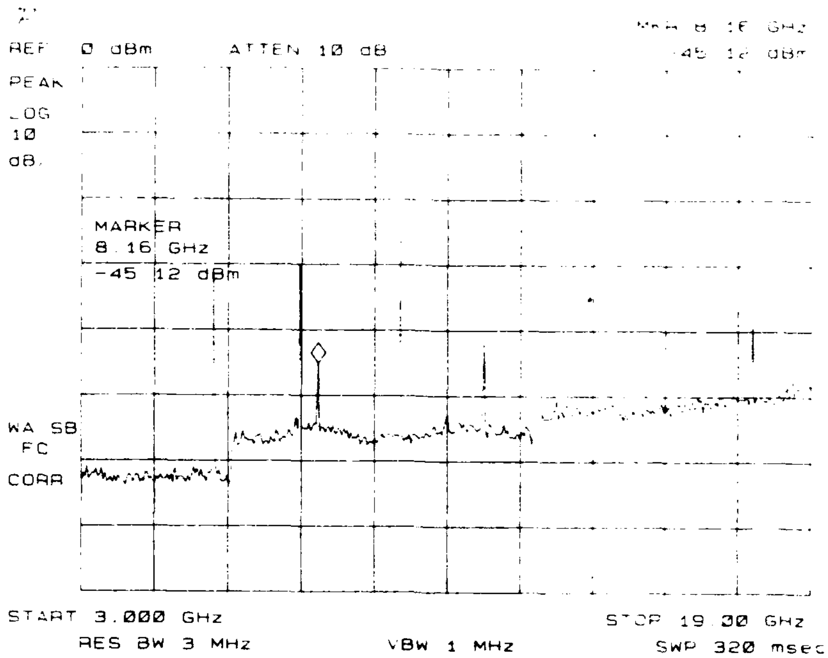

Fig. 2.7 Power Spectrum of the SOM before tuning RF signal

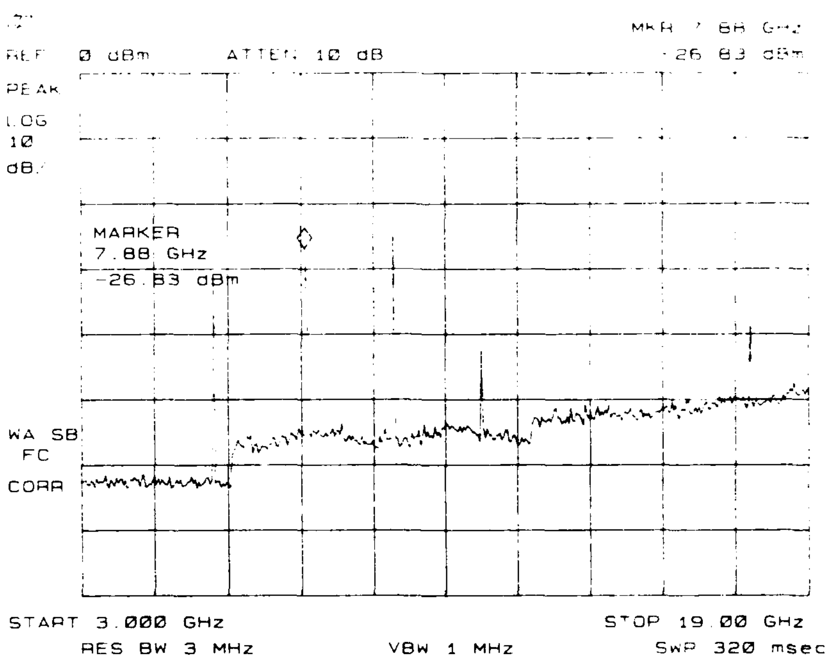

Fig. 2.8 Power Spectrum of the SOM after tuning RF signal

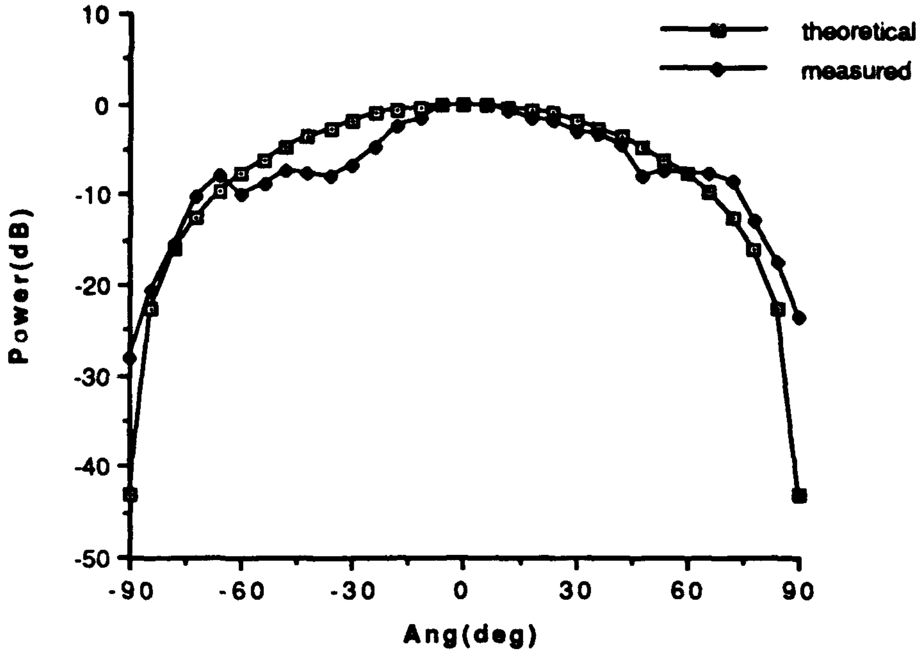

Fig. 2.9 The Receiving Antenna Pattern (H plane)

Chapter 3

Transmitting Part

3.1 Bandpass Filter and IF Amplifier Design

To pick only the second harmonic (8 GHz) of the first order mixing product and reflect other signals except 8 GHz back to the circuit reactively, a bandpass filter might be needed. Tchebyscheff methods were used for designing bandpass filter which is a parallel coupled type. The bandpass filter was designed for 7.7 GHz lower cutoff, 8.3 upper cutoff, 25 dB attenuation at 6 and 10 GHz, and 0.2 dB ripple. The third order bandpass filter was adopted to fulfill these requirements.

Since IF signal has a relatively low power, the amplifier is needed before transmitting that signal. To design the amplifier several items need to be considered [18]. The most important design considerations in a microwave transistor amplifier are stability, power gain, bandwidth, noise and dc requirements. A design usually starts with a set of specifications and the selection of the proper transistor. In this work, the Avantek's ATF-26884 FET which is a high performance GaAs Schottky-barrier-gate field effect transistor housed in a low cost-effective plastic package was chosen. This GaAs FET device has a nominal 0.3 micron gate length with a total gate periphery of 250 micron, 18 dBm the output power @ 1 dB gain compression, the typical value of gain about 11 dB and the optimum noise figure 2.2 dB at 12 GHz.

A systematic mathematical solution, aided by graphical methods (i.e. Smith Chart), is used to determine the transistor loading (i.e. the source and load reflection coefficients) for a particular stability and gain criteria An unconditionally stable transistor will not oscillate with any passive termination. On the other hand, a design using a potentially unstable transistor requires some analysis and careful

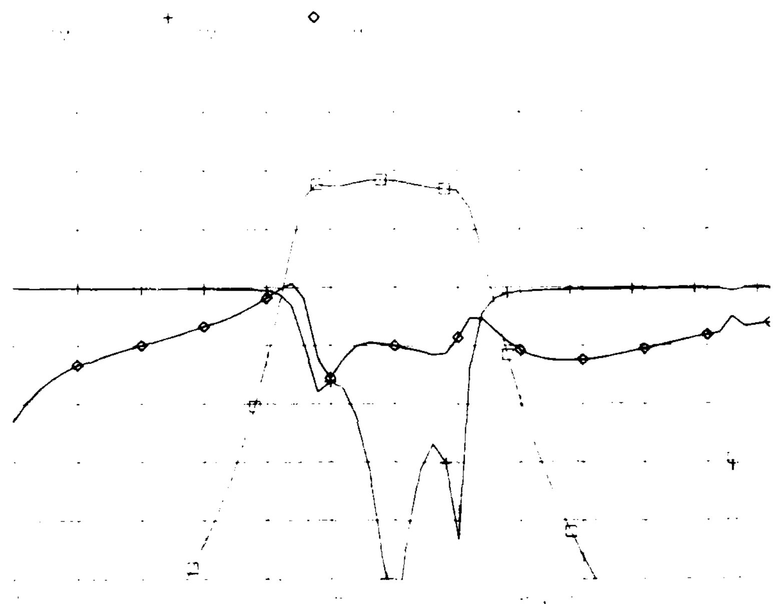

Fig. 3.1 The Simulation Result of the Transmitting Part

considerations so that thepassive terminations produce a stable amplifier.

As the IF amplifier, the low noise amplifier (LNA) is preferred. However, the vendor did not supply the noise parameter of the ATF-26884 which was chosen as an active source in this work. Hence, a high gain amplifier was designed . Even though the maximum stable gain of ATF-26884 is 10 dB at 10 GHz, the designed maximum gain was 9.5 dB considering stable operation. In whole circuit, 75 Ω characteristic impedance line was used for reduction of circuit dimension. The circuit file is attached in Appendix A and the simulation result including the bandpass filter and the amplifier is shown in Fig.3.1.

3.2 Patch Antenna Design

Originally the slot was adopted and tested but finally the rectangular patch is chosen for its simple geometry, easier fabrication compared with the slot. The patch antenna was among the earliest microstrip antennas developed, the structure and operation being related to that of resonators used in circuit design. The antenna consists of an isolated area of conductor on the upper surface of the microstrip substrate, with dimensions comparable to a half-wavelength, and is driven by a feed voltage applied between the conductor and the ground plane of the microstrip [20]. The configuration of the excitation of patch antennas which was adopted in this transponder is a microstrip transmission line etched upon the same circuit board, connected at the edge of the patch. This configuration has an advantage that circuit elements may be integrated upon the same printed circuit board, provided the extra space needed is available. A problem, however, is that the patch has a high-radiation impedance located at the edge. So the quarter-wave transformer was used for impedance matching between the circuit and the antenna [21].

The rectangular patch antenna is essentially a resonant microstrip with an electrical length of 1/2 the wavelength of the frequency to be transmitted or received. The input impedance of the patch antenna varies as a function of feed location. The edge of a 1/2 wavelength antenna has an input impedance of approximately 250 Ω which drops to zero Ω as the feed point is moved inboard to the center of the antenna. This provides a convenient means of impedance matching. Using EM simulator® two patch antennas were designed which has different resonant frequency; one is a receiving antenna (10 GHz) and another transmitting antenna (8 GHz). The dominant mode of these patch antennas is TE_{10}. The input impedances of each patch antenna are shown in Fig.3.2 and Fig.3.3. At the resonant frequency, the input impedance is about 250 Ω.

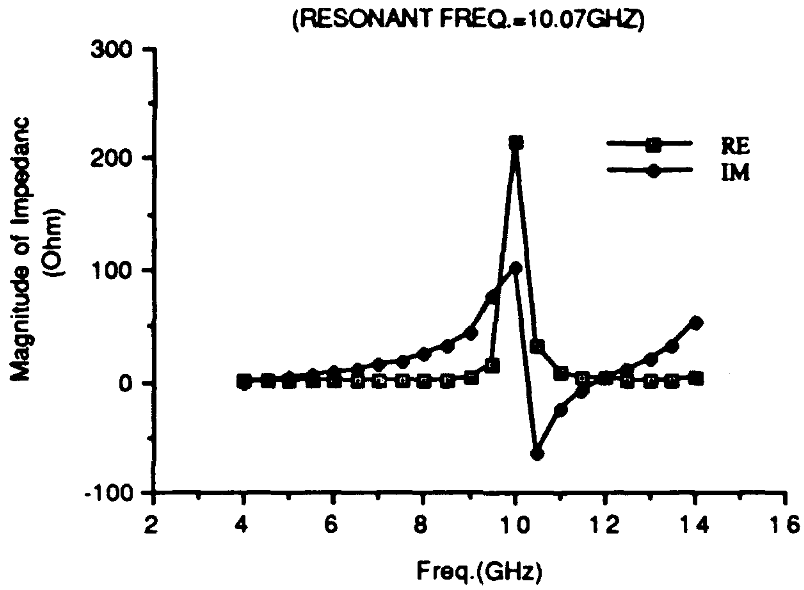

Fig. 3.2 Input Impedance of the Receiving Patch Antenna

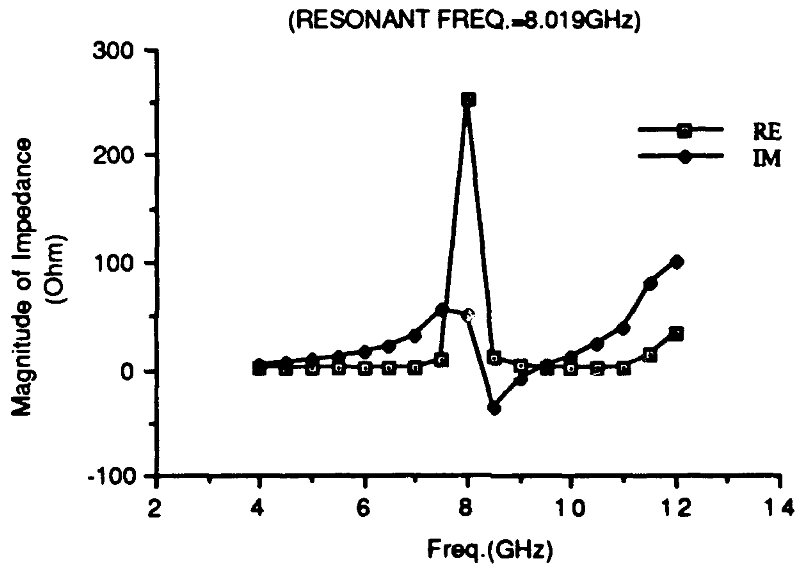

Fig. 3.3 Input Impedance of the Transmitting Patch Antenna

3.3 Measurement Results

The circuit of the transmitting part was fabricated as in Fig.3.4. The substrate and the FET were the same as the receiving part. Using the sweep oscillator 8 GHz signal fed into the bandpass filter through SMA connector. This signal was amplified by the IF amplifier and transmitted through the patch antenna to free space. The X-band standard gain horn which was located at 3 ft from the circuit was used to detect the signal and the signal was measured by the spectrum analyzer.

Fig.3.5 and Fig.3.6 show the measured signals transmitted by the circuit and the gain of the IF amplifier was 8.61 dB. Comparing the designed gain (9.5 dB) with this measured gain, the design of the amplifier was confirmed. If the higher gain is needed in this transmitting part, multi-stage amplifiers can be used. The measured transmitting antenna pattern in H plane was shown in Fig.3.7. Since there is single patch antenna, the theoretical antenna pattern has no side lobes [20]. The measured pattern was well matched to the theoretical one.

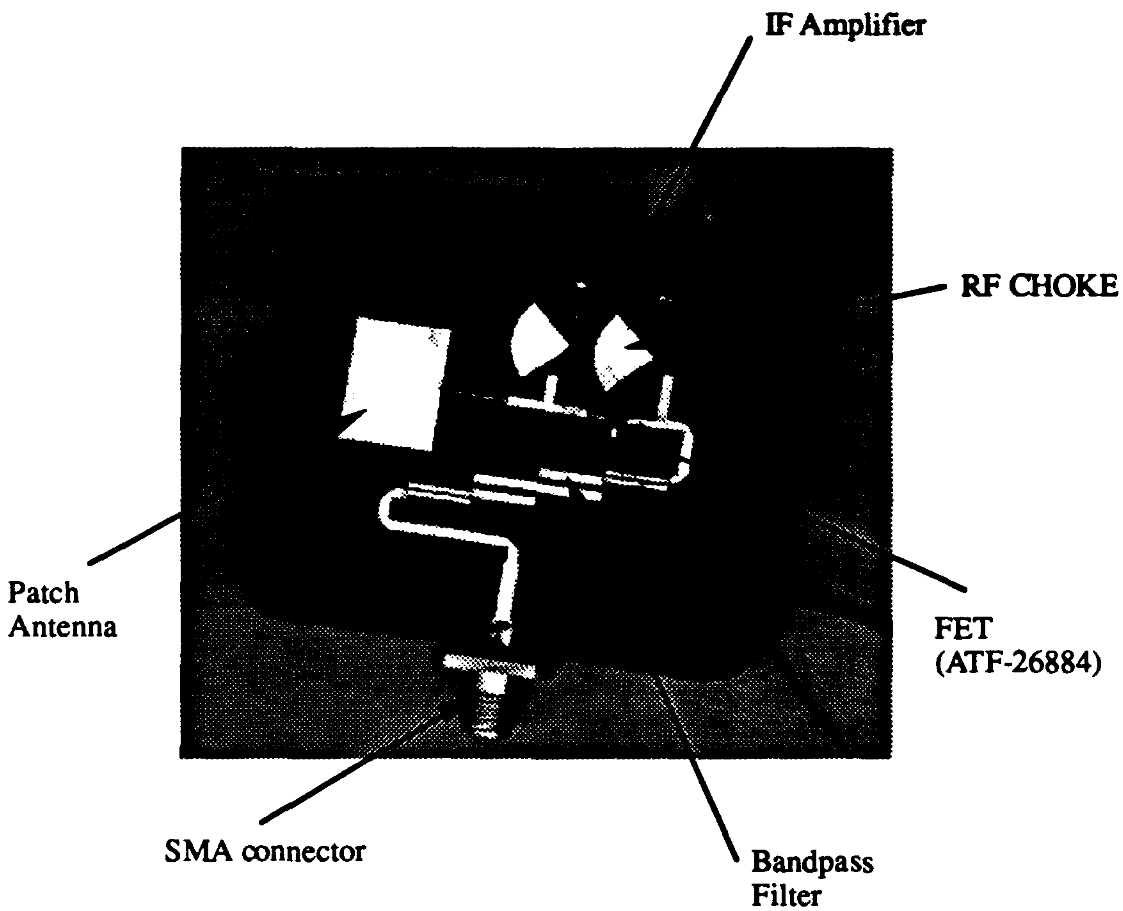

Fig. 3.4 Transmitting Part of Transponder

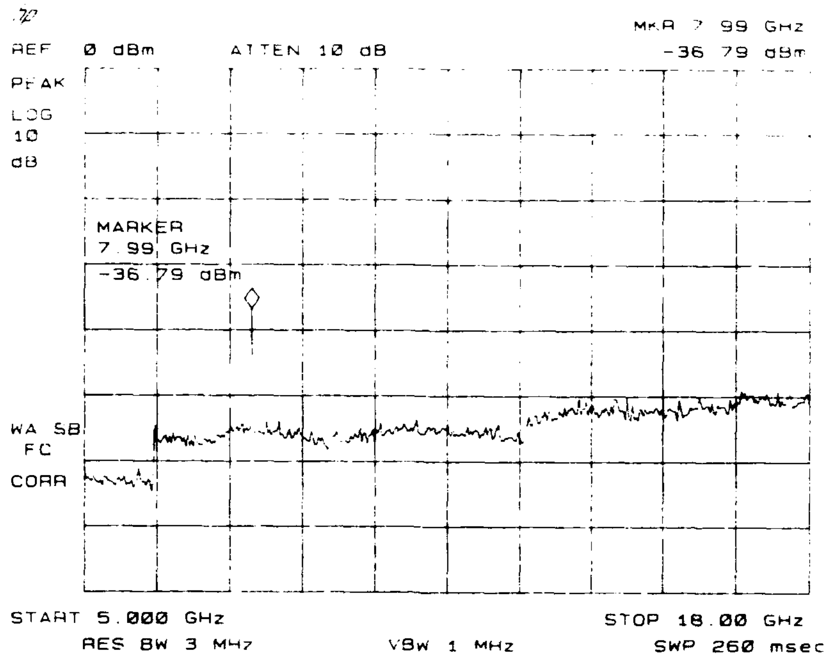

Fig. 3.5 Measured Signal from the Transmitting Part without amplification

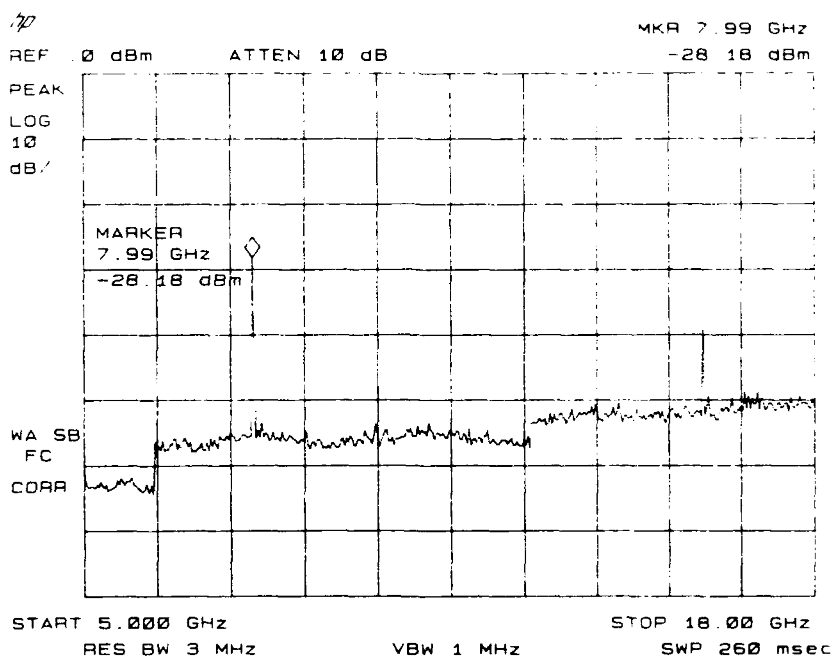

Fig. 3.6 Measured Signal from the Transmitting Part with amplification

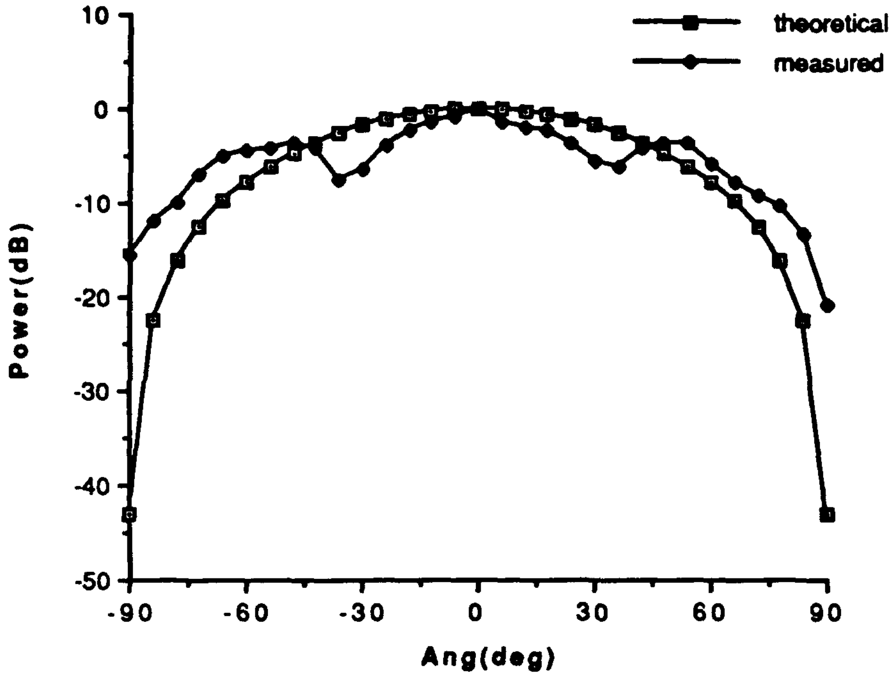

Fig. 3.7 The Transmitting Antenna Pattern (H plane)

Chapter 4

Transponder Using FETs and Patch Antenna

4.1 Circuit Design, Simulation, and Fabrication

The entire circuit of the transponder was designed and fabricated as shown in Fig.4.1. The dimension is 2.5 inch by 2.5 inch comparable to the usual credit card size. The concept of designing this transponder corresponds to that of the receiving part in Chapter 2. With the input impedance of both the transmitting part and the receiving patch antenna the one-port negative oscillator was designed as a whole circuit. Similarly, as already mentioned in Chapter 2, determining the location of the reference plane seemed not to be critical. The substrate and the FET were the same as the receiving and transmitting part.

In Chapter 2, 8 GHz which is the higher order mode, i.e. either the second harmonic of 4 GHz or the mixing product from 10 GHz RF and the third harmonic oscillation signal (6 GHz) was chosen as the IF signal. It was found that the IF signal level was too small to be detected. It could be worse if the entire circuit of the transponder is made for transmitting such higher order mixing products. The reason is that after cascading the SOM and the amplifier with patch antennas it is difficult to tell the conditions to support the recirculation of 4 GHz and have strong third harmonic of the oscillation signal enough to generate 8 GHz. Those conditions cannot be guaranteed in every circuits since one has to deal with the device which has nonlinear characteristics. Therefore, to transmit the signal over a certain distance which has large power level, the RF signal was changed from 10 GHz to 14 GHz. In this case, 8 GHz becomes the first order mixing product which has more power rather than any other mixing products.

The patch antenna was redesigned for 14 GHz of which the input impedance was 230 Ω. As mentioned already in Chapter 1 and 2, designing the SOM including antennas, the bandpass filter, and the IF amplifier was the most crucial thing. And one needs to consider the cascading the each component without deteriorating each function simultaneously. Since the load conditions of each component can be changed after cascading the components the impedance matching between the components should be taken care of. The program of the transponder is attached in Appendix B.

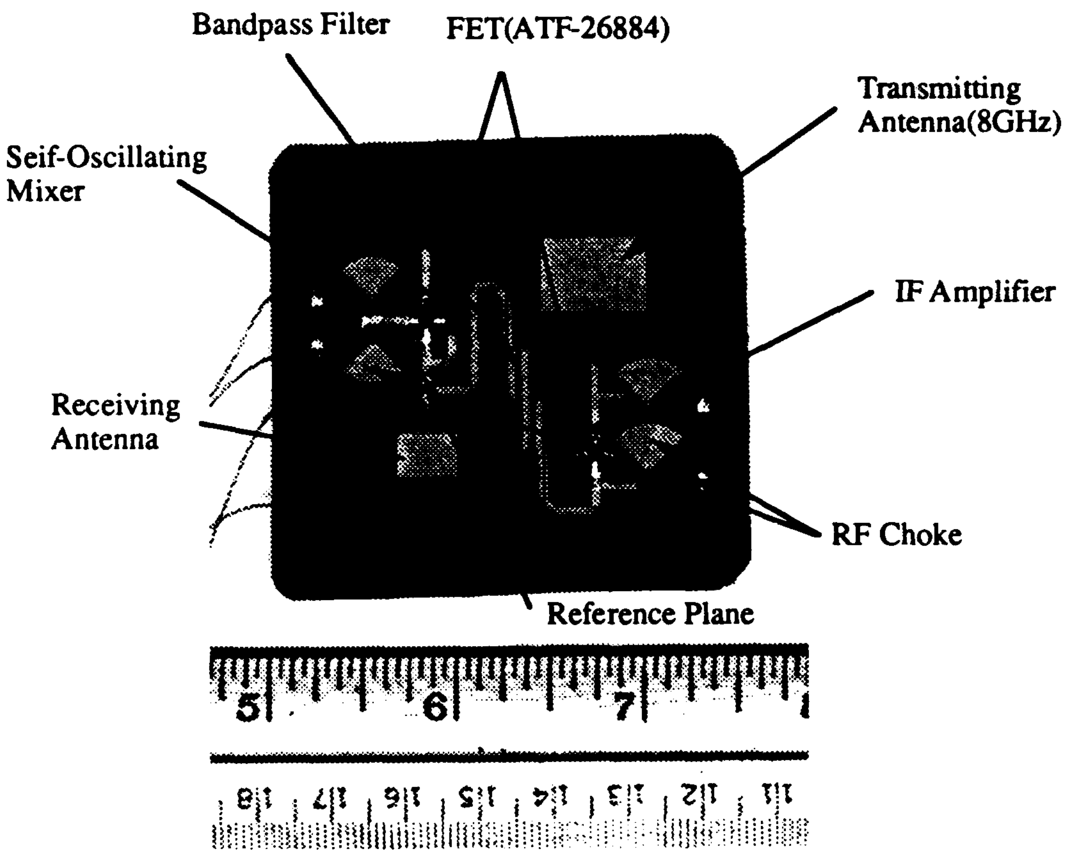

Fig. 4.1 Quasi-Optical Transponder

4.2 Measurement Result

Two different experimental setups were adopted to obtain the data from the transponder. The first experimental setup is shown in Fig.4.2. Only one horn antenna was used to transmit RF signal and receive IF signal. This horn antenna is an X-band standard gain horn. Using the computer-controlled step motor, the circuit was rotated to obtain the radiation pattern. Since the receiving signal level was small, the distance between the horn antenna and the transponder was set at 1 ft. The microwave amplifier was used to increase the RF power right after the sweep oscillator. The output power of RF signal at the output port of the microwave amplifier was 25 dBm. To separate the RF and LO signals the circulator was used.

Since the design concept of the transponder and the receiving part in Chapter 2 were the same as described in section 4.1, designing an one-port negative resistance oscillator was the first step. The designed oscillation frequency was 6.7 GHz. The simulation result was similar to Fig.2.6. The measured oscillation frequency was 6.66 GHz as shown in Fig.4.3 which is very close to the designed frequency. It was confirmed that this 6.66 GHz (LO signal) was transmitted through the 8 GHz patch antenna.

RF was at 14.69 GHz which was tuned to set the IF signal around 8 GHz since the LO signal was generated at 6.66 GHz. As seen in Fig.4.3 there were 4.89 GHz, 9.79 GHz, and 19.59 GHz signals located with the same distance from each other. These signals were produced by the sweep oscillator and detected by the spectrum analyzer via the circulator. The second and third harmonics of LO signal were also observed.

The measured amplifier gain was 5.33 dB. Comparing this with the amplifier gain of the transmitting part (8.61 dB) the gain was decreased. This may be due to the change of the load condition of the amplifier. After cascading the components, the source and load impedances as seen from the amplifier were changed slightly.

Hence, the performance of the amplifier was changed, i.e. somehow deteriorated.

The bias voltages of the FET were for the first FET in oscillator stage I_{DS}=4.6 mA, V_{DS}=2.96 V, and V_{GS}=-1.99 V and for the second FET in amplifier stage I_{DS}=10.0 mA, V_{DS}=3.04 V, and V_{GS}=-0.99 V. The detected power of IF signal was -41.29 dBm at 1 ft distance. When the distance was increased to 1.5 ft, the IF signal disappeared.

The second experimental setup is shown in Fig.4.4. In this setup, two horn antennas were used. One is a higher gain standard horn (gain = 24.7 dB at 10 GHz) to receive IF signal. The other is the same standard gain horn (gain = 16.4 dB at 10 GHz) as used in Fig.4.2 to transmit RF signal. The measured power spectrum is shown in Fig.4.5. Since the circulator was not used in this setup, there were no other signals produced except for LO, IF, and RF signals and no loss of RF signal caused

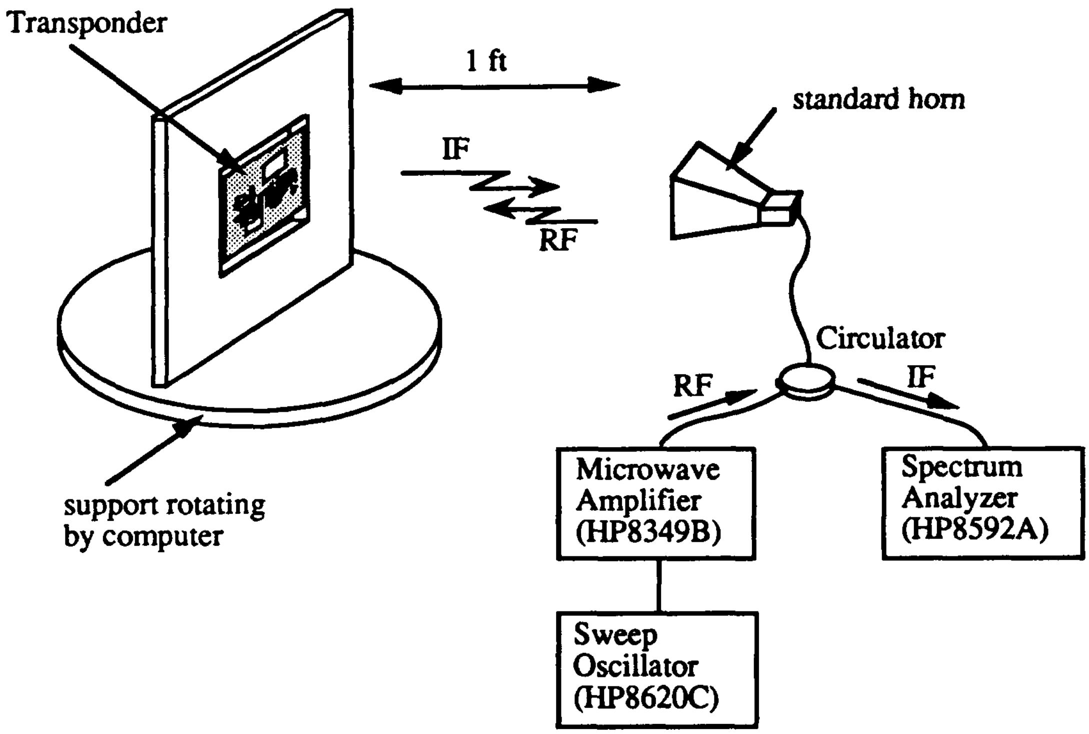

Fig. 4.2 Experimental Setup I

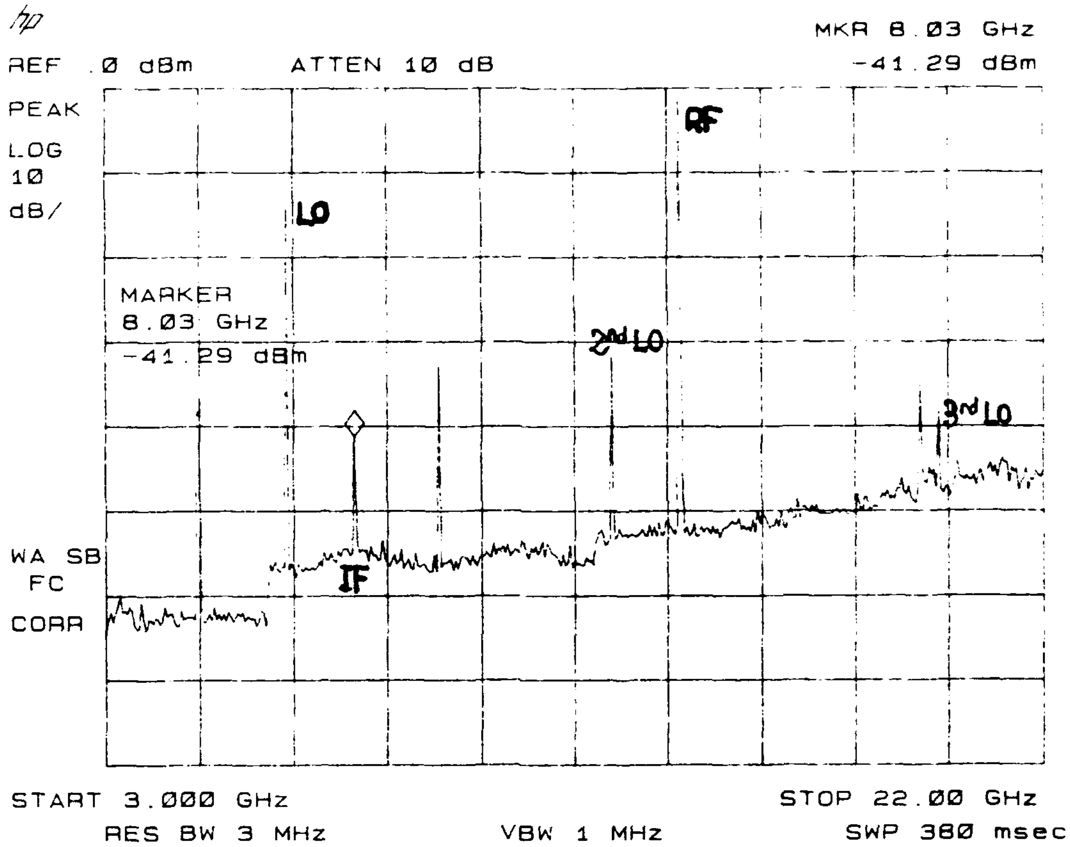

Fig. 4.3 Power Spectrum of the Transponder

by direct path to spectrum analyzer via the circulator. The oscillation frequency was generated at 5.18 GHz with different bias condition compared with the first setup (I_{DS}=2.0 mA, V_{DS}=1.391 V, and V_{GS}=-1.284 V). RF signal was at 12.22 GHz to set the LO signal be at around 7 GHz. The reason was that the transmitted power of around 7 GHz was higher than that of 8 GHz. This might be come from the changes of the bandpass filter characteristics. Since the higher gain horn was used and higher RF power was radiated, the IF signal could be detected over 2 ft distance. The measured power versus distance is shown in Fig.4.6. The result was improved compared with the measurement result of the first experimental setup. The IF power was detected until 6 ft. The power falls off proportional to one over distance square.

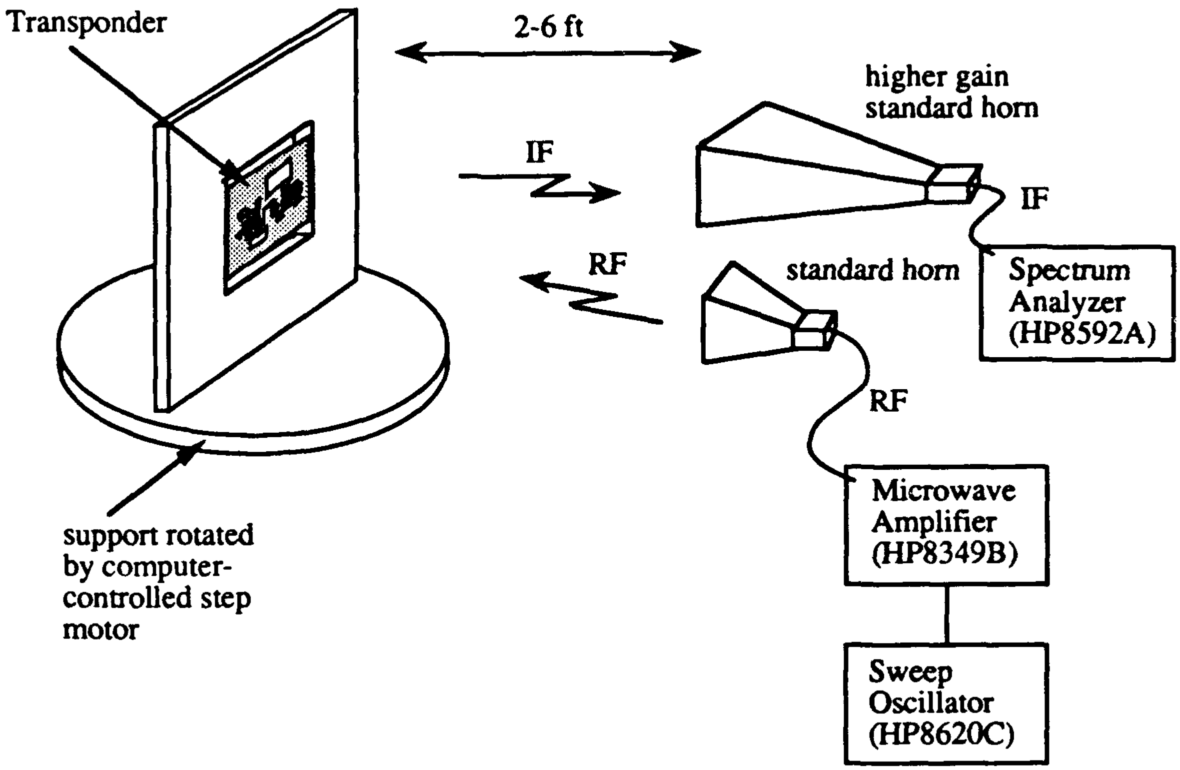

Fig. 4.4 Experimental Setup II

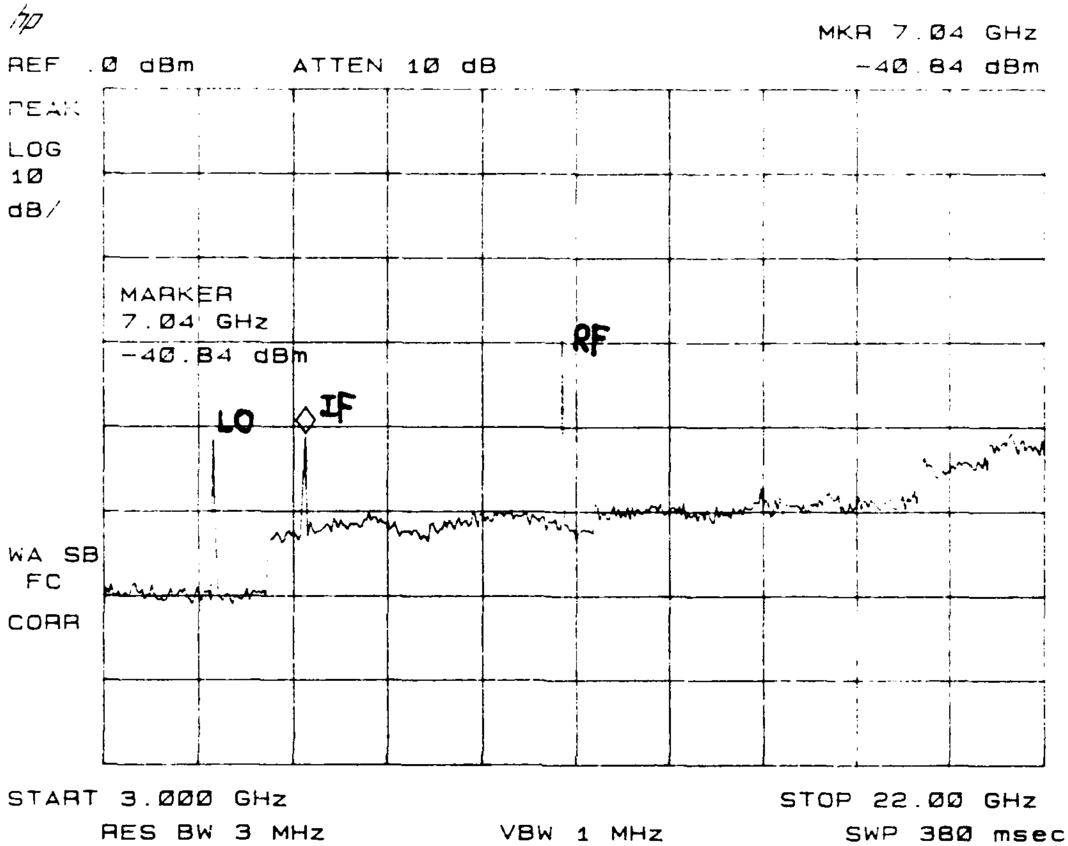

Fig. 4.5 Power Spectrum of the Transponder

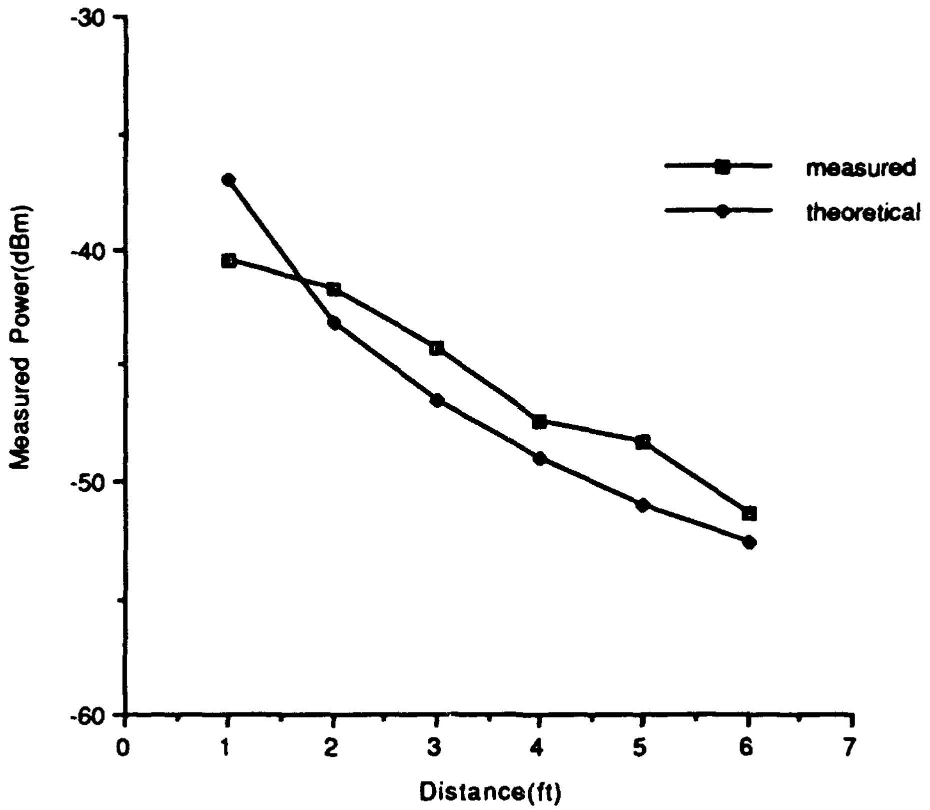

Fig. 4.6 Measured IF power versus Distance

Chapter 5

Conclusion

As presented through this thesis the quasi-optical transponder is designed and successfully demonstrated. To confirm the function of the transponder step by step, the transponder is divided into two parts and each one is designed and tested. The first part is the receiving part and the second is the transmitting part.

For the receiving part, the self-oscillating mixer (SOM) is used to generate local oscillation (LO) signal as well as intermediate frequency (IF) by mixing LO with RF. The RF signal is received through the patch antenna. To make only the IF signal (8 GHz) pass through,. the bandpass filter is adopted. As a simple receiver front end, this receiving part in itself has many applications.

For the transmitting part, the amplifier is used to increase the IF power. For stable operation, the unconditionally stable amplifier is designed. This transmitting part has potential to increase the gain if the multi-stage amplifier is employed. To confirm the cascade to the receiving part, the bandpass filter treated as the receiving part is included in this part. After each part is demonstrated successfully, the entire circuit of the transponder is made.

Since the concepts of active antenna and the SOM are used in designing this transponder it could be small and compact. The transponer is expected to be realized for many microwave and millimeter-wave applications: mobile communication systems, global positioning systems, and contactless ID systems. In the prototype circuits, the design of the SOM is carried out by the small signal analysis. Even though both the oscillator and the mixer are large signal devices the small signal analysis shows pretty accurate results.

For the future research suggestions, a large signal analysis will be needed which is not done in this thesis. For MMIC circuits where mechanical tweaking and tuning is not possible, a large signal analysis should be carried out. To increase the IF power of the transponder multi-stage IF amplifier can be used. Practically, considering the circuit size two-stage LNA (low noise amplifier) which has around 20 dB gain using HEMTs will be preferred.

Appendix A

Circuit File of IF Amplifier with Bandpass Filter

```
DIM                                    ! Default units in this circuit file
    FREQ      GHZ
    LNG       MIL                      ! 1 mm = 40 mil
    RES       OH
    CAP       PF
    IND       NH
    ANG       DEG

VAR
    WBIAS=8                            ! width of bias line
    LBIAS=260                          ! length of bias line
    WD=45.5                            ! width of transmission line of 75 Ω
    WI1=23                             ! <= bandpass filter parameters
    LL1=271
    SP1=9
    WI2=42
    SP2=27
    LL2=269
    LCON=20                            ! <= until here

    LLG1=50
    LLGB#35     43     55              ! gate length of the amplifier
    LLG2=28
    LOC1#190    198    210             ! open stub length of gate
    LLDB#20     30     45              ! drain length of the amplifier
    LLD1=155
    LOC2#115    125    135             ! open stub length of drain
    LLD2=150
    WDT1=65.4                     ! width of transition line from 75Ω to 50Ω
    WDT2=9.37                     ! width of transition line from 75Ω to 258Ω
                                   (input impedance of patch antenna at 8GHz)
```

```
CKT
   MSUB ER=2.33 H=31 T=1.4 RHO=1.0 RGH=0      ! RT/duroid substrate
! <== connect to bandpass filter
   ! SHOR     24
   ! RES      24     25     R=50
   ! MLIN     25     26     W=90  L=150           ! 50Ω line to SMA
   ! MSTEP    26     27     W1=90       W2^WDT1
   ! MLIN     27     28     W^WDT1      L=263.76
   ! MSTEP    28     29     Wl^WDT1     W2^WD
   MLIN       29     51     W^WD        L=50
   MBEND3     52     51     W^WD
   MLIN       52     53     W^WD        L=500
   MBEND3     53     54     W^WD
   MLIN       54     55     W^WD        L=80
   MBEND3     55     56     W^WD

   MLIN       56     62     W^WD        L=50
   MSTEP      62     63     Wl^WD       W2^WI1
   MCLIN      63 64 65 66   W^WI1  S^SP1  L^LL1      ! first coupled line
                            Wl^WD W2=0 W3^WI2 W4=0
   MSTEP      65     67     Wl^WI1      W2^WI2
   MCLIN      67 68 69 70   W^WI2  S^SP2  L^LL2      ! second coupled line
                            Wl^WI1 W2=0 W3^WI2 W4=0
   MSTEP      69     71     Wl^WI2      W2^WI2
   MCLIN      71 72 73 74   W^WI2  S^SP2  L^LL2      ! third coupled line
                            Wl^WI2 W2=0 W3^WI1 W4=0
   MSTEP      73     75     Wl^WI2      W2^WI1
   MCLIN      75 76 77 78   W^WI1  S^SP1  L^LL1      ! fourth coupled line
                            Wl^WI2 W2=0 W3^WD W4=0
   MSTEP      77     79     Wl^WI1      W2^WD
   MLIN       79     80     W^WD        L=20

   MBEND3     81     80     W^WD
   MLIN       81     82     W^WD        L=180
   MBEND3     83     82     W^WD
```

```
    MLIN       83    84       W^WD        L^LCON
! <== bandpass filter component

! ==================== IF Amplifier ====================
    S2PA       501  502 0     ATF2-43V.S2P              ! FET(ATF26884)
    ! MTEE     901  902 903   W1=130  W2=130  W3=140    ! space for FET
    ! MGAP     501   901      W^WD        S=10          ! used in MICAD
    ! MGAP     902   502      W^WD        S=10
    ! MGAP     903   503      W=10        S=10
! ------------------------------------ gate ------------------------------------
    MLIN       84    511      W^WD        L^LLG1
    MTEE       512 511 513    Wl^WD   W2^WD   W3^WD
    MLOC       513            W^WD        L^LOC1        ! matching stub
    MLIN       512   514      W^WD        L^LLGB
    MTEE       551 514 552    Wl^WD   W2^WD   W3^WBIAS       ! connect to
    MLIN       552   553      W^WBIAS     L^LBIAS            ! bias line
    MTEE       553 554 555    Wl^WBIAS W2^WBIAS W3^WBIAS
    MRSTUB     555            WI^WBIAS  L=250  ANG=90
    MLIN       554   556      W^WBIAS     L^LBIAS
    MSTEP      556   557      Wl^WBIAS    W2=60

    MLIN       557   558      W=60        L=60
    MLIN       551   501      W^WD        L^LLG2
! ------------------------------------ drain ------------------------------------
    MLIN       502   520      W^WD        L^LLD1
    MTEE       561 520 562    Wl^WD  W2^WD  W3^WBIAS         ! connect to
    MLIN       562   563      W^WBIAS     L^LBIAS            ! bias line
    MTEE       563 564 565    Wl^WBIAS W2^WBIAS W3^WBIAS
    MRSTUB     565            WI^WBIAS L=250  ANG=90
    MLIN       564   566      W^WBIAS     L^LBIAS
    MSTEP      566   567      Wl^WBIAS    W2=60

    MLIN       567   568      W=60        L=60
    MLIN       561   521      W^WD        L^LLDB
    MTEE       522 521 523    Wl^WD   W2^WD   W3^WD
```

```
    MLOC      523          W^WD       L^LOC2        ! matching stub
    MLIN      522    524   W^WD       L^LLD2
    ! RES     524    525   R=75
    ! SHOR    525
    ! MSTEP   524    525   W1^WD      W2^WDT2
    ! MLIN    525    526   W^WDT2     L=274.80
    ! MSTEP   526    527   W1^WDT2    W2=567
    ! S1PA    527      0   PATCH8.Z1P               ! 8GHz patch antenna
    ! MLIN    527    528   W=567      L=403
    ! RES     528    529   R=258
    ! SHOR    529
!
    DEF2P     29     524   AMP

TERM
    Z0=75                             ! input and output load impedance

FREQ
    SWEEP     5      11    0.1
OUT
    AMP       RE[Z1]       GR1
    AMP       IM[Z1]       GR1
    AMP       DB[S21]      GR1        ! gain of the amplifier
    AMP       DB[S11]      GR1        ! return loss of the amplifier
    AMP       DB[S22]      GR1
    AMP       VSWR1        GR2        ! input standing wave ratio
    AMP       VSWR2        GR2        ! output standing wave ratio

GRID
    RANGE      5     11    0.5
    GR1      -25     20    5
    GR2        0     10
OPT
    RANGE     7.8    8.2
    AMP       DB[S21] = 11            ! gain for optimization
```

Appendix B

Circuit File of the Quasi-Optical Transponder Using FETs and Patch Antennas

```
DIM                                         ! Default units in this circuit file
    FREQ        GHZ
    LNG         MIL                         ! 1 mm = 40 mil
    RES         OH
    CAP         PF
    IND         NH
    ANG         DEG

VAR
    WBIAS=8                                 ! width of bias line
    LBIAS=260                               ! length of bias line
    WGE=45.5
    LOCG#100    265     500                 ! gate length of the oscillator
    WFS=45.5                                ! width of source element
    LFS#100     260     650                 ! length of source
    WD=45.5                                 ! 75Ω transmission line
    LD1#30      90      150                 ! drain length of the oscillator
    LD2#50      51      150                 ! drain length of the oscillator
    LD3=50
    LOCD1#30    110     300                 ! length of drain open stub
    WI1=23                                  ! <= bandpass filter parameters
    LL1=271
    SP1= 9
    WI2=42
    SP2=27
    LL2=269
    LCON=20                                 ! <= until here
    LLG1=50                                 ! <= IF amplifier parameters
    LLGB=43
    LLG2=28
```

```
    LOCl=198                                    ! matching stub length of gate part
    LLDB=30
    LLDl=155
    LOC2=125                               ! matching stub length of drain part
    LLD2=150                               ! <= until here
    WDT1=11.65                             ! width of transition line from 75Ω to 230Ω
                                          (input impedance of patch antenna at 14GHz)
    WDT2=9.37                              ! width of transition line from 75Ω to 258Ω
                                           (input impedance of patch antenna at 8GHz)
CKT
    MSUB ER=2.33 H=31 T=1.4 RHO=1.0 RGH=0     ! RT/duroid substrate

! =========== Self-Oscillating Mixer with Band-pass filter ===============
    S2PA        1  2  3       ATF2-43V.S2P        ! FET (ATF26884)
    ! MTEE      202 201 203   Wl=130  W2=130  W3=140  ! space for FET
    ! MGAP       1     201    W^WGE       S=10        ! used in MICAD
    ! MGAP      202      2    W^WD        S=10
    ! MGAP      203      3    W=10        S=10
    MLOC        10            W^WGE       L^LOCG      ! gate resonator
    MTEE        11   10  12   Wl^WGE  W2^WGE  W3^WBIAS
    MLIN        12      101   W^WBIAS     L^LBIAS     ! bias line
    MTEE        102  101  103 Wl^WBIAS  W2^WBIAS  W3^WBIAS
    MLIN        102     104   W^WBIAS     L^LBIAS
    MRSTUB      103           Wl^WBIAS  L=220  ANG=90
    MSTEP       104     105   Wl^WBIAS    W2=60
    MLIN        105     106   W=60        L=60        ! space for solder

    MLIN         11       1   W^WGE       L=30
! ------------------------------------ source ------------------------------------
    MLSC          3           W^WFS       L^LFS       ! source element
! ------------------------------------ drain -------------------------------------
    MLIN        2       21    W^WD        L=30
    MTEE        22 21 23      Wl^WD  W2^WD  W3^WBIAS
    MLIN        23      301   W^WBIAS     L^LBIAS     ! bias line
    MTEE        301 302 303   Wl^WBIAS  W2^WBIAS  W3^WBIAS
```

```
MLIN      302   304    W^WBIAS    L^LBIAS
MRSTUB    303          W1^WBIAS  L=220  ANG=90
MSTEP     304   305    W1^WBIAS   W2=60
MLIN      305   306    W=60       L=60

MLIN       22    24    W^WD       L^LD1
MTEE       24  25  26  W1^WD  W2^WD  W3^WD
MLIN       26   427    W^WD       L=80
MBEND3    428   427    W^WD
MLOC      428          W^WD       L^LOCD1     ! matching stub
MLIN       25   500    W^WD       L^LD2
UNIT      500   501                           ! reference plane for
UNIT      501   502                           ! designing the oscillator
MLIN      502    27    W^WD       L=50
MTEE       27  28  29  W1^WD  W2^WD  W3^WD

! <== connect to bandpass filter
MLIN       29    51    W^WD       L=200
MBEND3     52    51    W^WD
MLIN       52    53    W^WD       L=500
MBEND3     53    54    W^WD
MLIN       54    55    W^WD       L=80
MBEND3     55    56    W^WD

MLIN       56    62    W^WD       L=50
MSTEP      62    63    W1^WD      W2^WI1
MCLIN     63 64 65 66  W^WI1  S^SP1  L^LL1     ! first coupled line
                       W1^WD W2=0 W3^WI2 W4=0
MSTEP      65    67    W1^WI1     W2^WI2
MCLIN     67 68 69 70  W^WI2  S^SP2  L^LL2     ! second coupled line
                       W1^WI1 W2=0 W3^WI2 W4=0
MSTEP      69    71    W1^WI2     W2^WI2
MCLIN     71 72 73 74  W^WI2  S^SP2  L^LL2     ! third coupled line
                       W1^WI2 W2=0 W3^WI1 W4=0
MSTEP      73    75    W1^WI2     W2^WI1
```

```
    MCLIN     75 76 77 78   W^WI1   S^SP1   L^LL1        ! fourth coupled line
                            W1^WI2 W2=0 W3^WD W4=0
    MSTEP     77     79     W1^WI1      W2^WD
    MLIN      79     80     W^WD        L=20
    MBEND3    81     80     W^WD
    MLIN 81   82            W^WD        L=180
    MBEND3    83     82     W^WD
    MLIN      83     84     W^WD        L^LCON
! <== end of bandpass filter

    MLIN      28     30     W^WD        L^LD3
    MSTEP     30     31     W1^WD       W2^WDT1      ! transition from 75Ω
    MLIN      31     32     W^WDT1      L=156.12     ! to 230Ω
    ! MSTEP   32     33     W1^WDT1     W2=326
    ! MLIN    33     34     W=326       L=231
    S1PA      32      0     PATCH14.Z1P        ! 14GHz patch antenna

! ====================== IF amplifier ======================
    S2PB      601  602  0   ATF2-43V.S2P
    ! MTEE    901 902 903   W1=130  W2=130  W3=140
    ! MGAP    601    901    W^WD        S=10
    ! MGAP    902    602    W^WD        S=10
    ! MGAP    903    603    W=10  S=10
! ---------------------------------------- gate ----------------------------------------
    MLIN       84    611    W^WD        L^LLG1
    MTEE      612 611 613   W1^WD  W2^WD  W3^WD
    MLOC      613           W^WD        L^LOC1
    MLIN      612    614    W^WD        L^LLGB
    MTEE      651 614 652   W1^WD  W2^WD  W3^WBIAS
    MLIN      652    653    W^WBIAS     L=255        ! bias line
    MTEE      653 654 655   W1^WBIAS  W2^WBIAS  W3^WBIAS
    MRSTUB    655           W1^WBIAS  L=255  ANG=90
    MLIN      654    656    W^WBIAS     L=255
    MSTEP     656    657    W1^WBIAS    W2=60
    MLIN      657    658    W=60        L=60
```

```
	MLIN	651	601	W^WD	L^LLG2
! ---------------------------------------- drain ----------------------------------------
	MLIN	602	620	W^WD	L^LLD1
	MTEE	661 620 662	W1^WD  W2^WD  W3^WBIAS
	MLIN	662	663	W^WBIAS	L^LBIAS
	MTEE	663 664 665	W1^WBIAS  W2^WBIAS  W3^WBIAS
	MRSTUB	665		W1^WBIAS L=255  ANG=90
	MLIN	664	666	W^WBIAS	L=255
	MSTEP	666	667	W1^WBIAS	W2=60
	MLIN	667	668	W=60	L=60

	MLIN	661	621	W^WD	L^LLDB
	MTEE	622 621 623	W1^WD  W2^WD  W3^WD
	MLOC	623		W^WD	L^LOC2
	MLIN	622	624	W^WD	L^LLD2
	! RES	624	625	R=75
	! SHOR	625

	MSTEP	624	625	W1^WD	W2^WDT2
	MLIN	625	626	W^WDT2	L=274.80
	! MSTEP	626	627	W1^WDT2	W2=567
	! MLIN	627	628	W=567	L=403
	! RES	628	629	R=258
	! SHOR	629

	DEFIP	501		TRFD

FREQ
	SWEEP	4	10	0.1

OUT
	TRFD	RE[Z1]	GR1	! resistance of the oscillator
	TRFD	IM[Z1]	GR1	! reactance of the oscillator
	TRFD	DB[S11]	GR2
	TRFD	S11	SC2
```

```
    TRFD        DB[S11]      GR3

GRID
    RANGE         4     10    0.5
    GR1         -20     20      5
    GR2         -50     50     10
    GR3         -50      5      5

OPT
    RANGE       6.6    7.0
    TRFD        RE[Z1]=-6
    TRFD        IM[Z1]=0
```

Bibliography

[1] Y. Tajima, "GaAs FET Applications for Injection-Locked Oscilllator and Self-Oscillating Mixers", IEEE Trans. Microwave Theory Tech., vol. MTT-27, pp. 629-632, July 1979.

[2] V.D. Hwang and T. Itoh, "Quasi-Optical HEMT and MESFET Self-Oscillating Mixers", IEEE Trans. Microwave Theory Tech., vol. 36, pp. 1701-1705, Dec 1988.

[3] L. Yuan, J. Paul, and P. Yen, "140 GHZ quasi-optical planar mixer", 1982 MTT Int. Microwave Symp. Dig., pp. 374-375.

[4] K. Stephen, N. Camilleri, and T. Itoh, "A quasi-optical polarization diplexed balanced mixer for millimeter-wave applications", IEEE Trans. Microwave Theory Tech., vol. MTT-31, pp. 164-170, Feb. 1987.

[5] K. Stephen, and T. Itoh, "A planar quasi-optical sub-harmonically pumped mixer charcterized by isotropic conversion loss", IEEE Trans. Microwave Theory Tech., vol. MTT-32, pp. 97-102, Jan. 1984.

[6] S.Kawasaki and T. Itoh, "Quasi-Optical and Active Antenna Technology", the Conference Proceedings of MM-92, Brighton, UK, Oct. 1992, pp.311-316.

[7] S.A. Mass, *Nonlinear Microwave Circuits,* MA:Artech House, 1988.

[8] V.D. Hwang, T. Uwano, and T. Itoh, "Quasi-optical integrated antenna and receiver front end", IEEE Trans. Microwave Theory Tech., vol. MTT-36, pp. 80-85, Jan. 1988.

[9] S.A. Mass, *Microwave Mixers*, Artech House, Dedham, MA, 1986.

[10] R. Pucel, D. Masse, and R. Bera, "Performance of GaAs MESFET mixers at X band", IEEE Trans. Microwave Theory Tech., vol. MTT-24, pp. 351-360, June 1976.

[11] P. Bura and R. Dikshit, "FET mixers for communication satellite transponders", 1976 Int. Microwave Symp. Dig., pp. 90-92.

[12] G. Begemann, and A. Hecht, " The conversion gain and stability of MESFET gate mixers", 9th Europ. Microwave Conf. Proc. pp. 316-320, 1979.

[13] G. Tie, and C. Aitchison, "Noise figure and Associated Conversion gain of a microwave MESFET gate mixer", 13th Europ. Microwave Conf. Proc., pp.579-584, 1983.

[14] S. Maas, "Theory and Analysis of GaAs MESFET mixers", IEEE Trans. Microwave Theory and Tech., vol. MTT-32, pp. 1402-1406, Oct. 1984.

[15] C. Camacho-Penalosa, and C. Aitchison, "Analysis and design of MESFET gate mixers", IEEE Trans. Microwave Theory and Tech., vol. MTT-35, pp.643-652, July 1987.

[16] G. Begemann, and A. Jacob, "Conversion gain of MESFET drain mixers", Electron. Lett., vol. 15, pp. 567-568, Aug. 1979.

[17] S. Maas, "A GaAs MESFET mixer with very low intermodulation", IEEE Trans. Microwave Theory and Tech., vol. MTT-35, pp. 425-429, Apr. 1987.

[18] G. Gonzalez, *MICROWAVE TRANSISTOR AMPLIFIERS Analysis and Design*, Prentice-Hall, Englewood Cliffs, NJ, 1984

[19] K. Kurokawa, "Some Basic Characteristics of Broadband Negative Resistance Oscillator Circuits", The Bell System Technical Journal, July 1969.

[20] J. R. James, P. S. Hall, and C. Wood, *MICROSTRIP ANTENNA* Theory and Design, Peter Peregrinus Ltd., Stevenge, UK and NY, 1981.

[21] J. Lin and T. Itoh, "Two-dimensional quasi-optical power-combining arrays using strongly coupled oscillators", submitted to the Mini-Special Issue on Space Terahertz Technology of IEEE Trans. Microwave Theory and Tech., April 1994.