# Millimeter-Wave Monolithic GaAs IMPATT VCO

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*Abstract*—**A monolithic voltage-controlled oscillator (VCO) has been constructed using a GaAs double-drift Read IMPATT as the active element and a similar diode biased below breakdown as the varactor. The chip produced 120 mW peak power over an electronically controlled tuning range between 47 and 48 GHz. A computer analysis based on characterized circuit parameters has been used to predict the performance of the chip.**

## I. Introduction

Millimeter-wave voltage-controlled IMPATT diode oscillators are useful in many applications. Electronic frequency tuning is often achieved by varying circuit reactance using a varactor diode. Because of differing design requirements, it is difficult to integrate two types of devices in the same monolithic circuit [1]. However, the similarity of doping profile for hyperabrupt varactors and GaAs Read IMPATT diodes suggests an approach to the development of a monolithic varactor-tuned oscillator [2]–[4].

In the present work, a double-drift Read IMPATT doping profile was used to test a monolithic VCO design. Although double-drift IMPATT's have a higher dc to RF conversion efficiency than single-drift IMPATT's, the low mobility of holes results in a high parasitic series resistance and reduces the device $Q$ when a double-drift IMPATT is used as a varactor. This effect will be seen in the test data which follow.

## II. Varactor Characteristics

Fig. 1(a) shows the profile of the doping concentration and the electric field within a single-drift Read IMPATT. For the IMPATT diode, the spike determines the avalanche zone and improves the dc to RF conversion. When the diode is biased below its breakdown voltage, it can also be used as a varactor [5].

At zero bias voltage, the depletion region is confined within the charge spikes. As the reverse bias voltage increases, the depletion width increases, resulting in decreasing capacitance. Parasitic series resistance results from undepleted n-type material.

For a double-drift Read IMPATT profile as used in the present experiments, additional parasitic series resistance occurs in the varactor. The double-drift Read profile is shown in Fig. 1(b). The junction lies between two doping

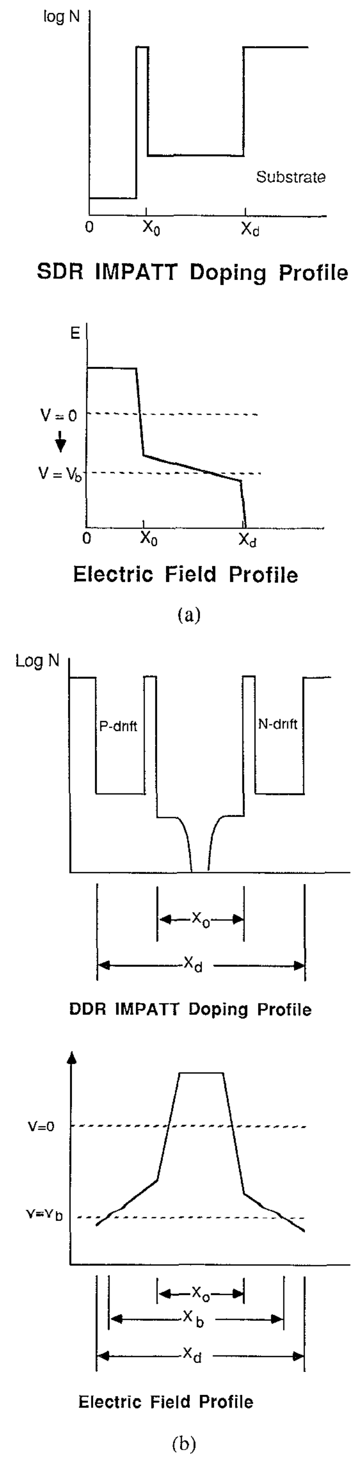


Fig. 1. Doping profile and electric field within an IMPATT diode. (a) Single-drift Read IMPATT. (b) Double-drift Read IMPATT.

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spikes, and drift zones extend into both the p and n IMPATT drift regions. The parasitic series resistance per micron of undepleted semiconductor is an order of magnitude greater for p material compared with n-type GaAs. Typically, the total charge in the p spike is somewhat less than that in the n spike. Therefore the p region is fully depleted prior to the n region. We gain little advantage from this since the capacitance is nearly saturated at the p side punch-through voltage.

With the above factors duly noted, we selected a double-drift doping profile which gave good dc to RF conversion efficiency in the IMPATT mode and strong capacitance variation with minimum loss as a function of applied varactor voltage.

## III. Circuit Design

The circuit design was based upon the monolithic IMPATT oscillator previously reported [6]. To incorporate the varactor into the oscillator, the following requirements were met:

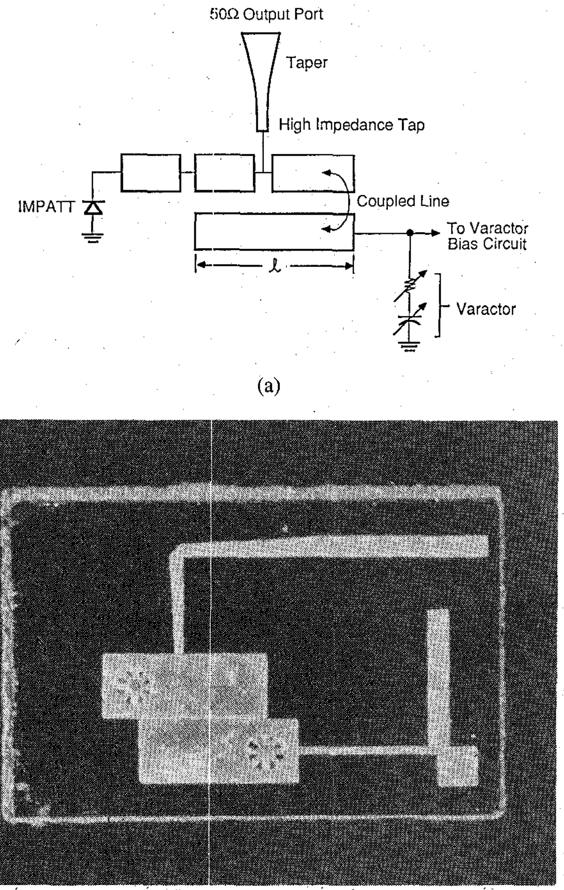
1) The varactor requires a separate dc bias voltage. Therefore, except for a common ground, the IMPATT and the varactor must be dc isolated.
2) The varactor must be RF coupled to the oscillator. Stronger coupling will give a greater frequency tuning range, but parasitic series resistance in the varactor will load the oscillator, thereby reducing the output power.
3) Device processing should be as simple as possible. Thin film resistors and MIM capacitors, although available, should be avoided.

A schematic drawing and photograph of a completed chip are shown in Fig. 2. The diodes are connected in shunt to the microstrip tuning elements by the air bridge leads seen in the photo [6]. Fig. 3 is a cross-sectional view. The diodes are positioned on a gold-plated via hole.

The oscillator incorporates an open-ended resonator with a high impedance tap for RF output coupling. A taper transforms the tap impedance from 90 to 50 Ω. The location and the impedance ratio between the tap and the resonator line determine the output coupling.

The resonator is a section of 25 Ω transmission line with total length over one-quarter wavelength to provide an inductive load impedance to the IMPATT. By means of the tap, dc bias is provided to the IMPATT. The bias tee is incorporated in the microstrip to waveguide transition so that a monolithic bias circuit for the IMPATT can be omitted in the present design.

A coupled line is used to establish the interaction between the IMPATT and the varactor. The varactor is connected in shunt to another 25 Ω microstrip line section, which is located alongside the resonator (forming a coupled line). The gap between the two lines is 10 μm. The remaining circuitry provides the varactor bias. Near the operation frequency, the bias circuit presents a high impedance to the varactor, with little effect on the VCO design.




(b)

Fig. 2. (a) Schematic drawing of the VCO circuit and (b) photograph.

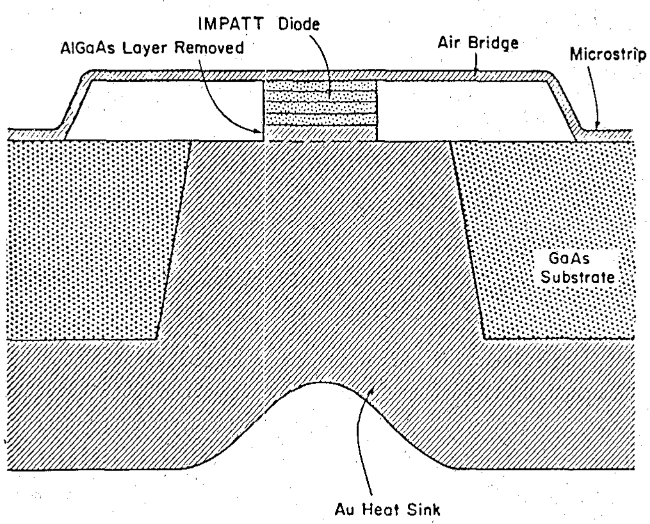



Fig. 3. Monolithic IMPATT diode structure.

An analysis of coupled lines appropriate to the present case can be found in the literature [7]. Fig. 4 shows the coupling structure. The input impedance can be expressed by

$$Z_{in} = Z_{11} - \frac{Z_{13}^2}{Z_{11} + Z_3}$$

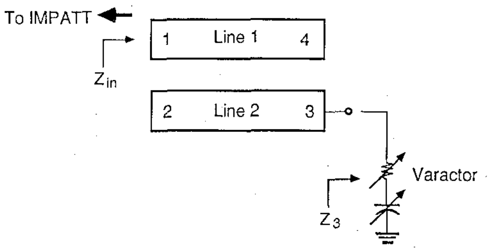



Fig. 4. Varactor tuning circuit.

where

$$Z_3 = R_s(V) + [j\omega C(V)]^{-1}$$

$$Z_{11} = \frac{Z_{0e} + Z_{0o}}{2} \coth \gamma l$$

$$Z_{13} = \frac{Z_{0e} - Z_{0o}}{2} \operatorname{csh} \gamma l.$$

$Z_{0e}$ and $Z_{0o}$ are the even- and odd-mode impedances of the coupled line. The quantity $\gamma = \alpha + j\beta$ is the complex propagation constant, assumed to be the same for both even and odd modes.

The highest sensitivity to varactor tuning occurs when the coupled line length is about a quarter wavelength. In that case, $Z_{11}$ will be zero. Also, the stronger the coupling, the larger $Z_{13}$ is.

The above discussion outlines the general features of our design approach. Since the assumption of equal propagation constants for both modes of coupled line is an approximation, detailed design was conducted with the aid of a computer. Measured circuit parameters [5] were used in the analysis.

## IV. Experimental Results

The monolithic VCO was fabricated on a double-drift Read profile GaAs IMPATT wafer. An AlGaAs stop etch layer is incorporated for via hole processing [6]. The diode mesa is defined first. The top contact is connected to the microstrip line circuit through an air bridge. Then the substrate is lapped to 100 $\mu$m. The via hole is opened by dry etching, which stops at the AlGaAs layer. A metal layer is plated over the entire back side to form the ground plane of the microstrip line. The via hole, which is the path for both electric signal and heat, is partially filled by gold plating. The chip measures 2 mm $\times$ 1.3 mm.

RF testing is done with a finline transition between WR-22 waveguide and microstrip line. The bias tee function is included in the transition, providing bias to the IMPATT diode. The insertion loss of the transition is in the range 0.5–0.8 dB over the entire waveguide band. Fig. 5 is a picture of the transitions.

The VCO was tested in pulse operation using a pulse width of 300 ns and a duty cycle of 6 percent. Fig. 6 is a typical result, showing the operation frequency and the output power as a function of the varactor bias voltage. A 1 GHz tuning range was achieved. The power dropped monotonically as the varactor bias voltage declined. Below 3.5 V oscillation is no longer obtained. This phenomenon strongly suggests that the series resistance of the varactor diode provides excessive loading to the IMPATT diode.

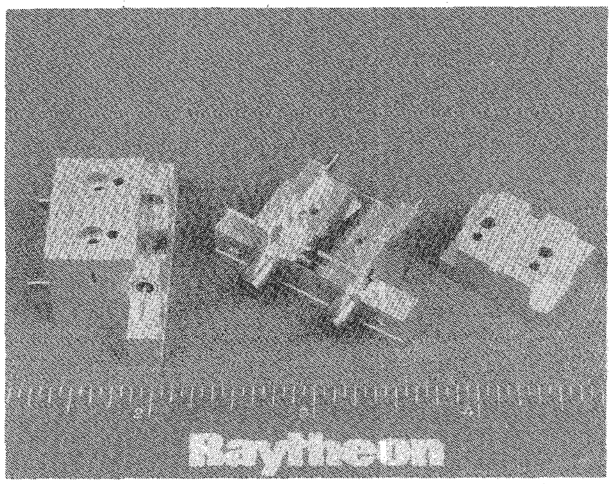



Fig. 5. Photograph of the waveguide (WR-22) to microstrip line transitions.

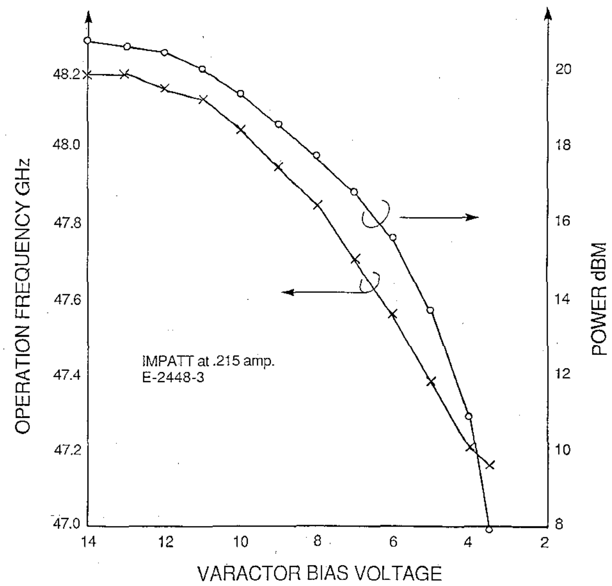



Fig. 6. Operation frequency and output power versus the varactor bias voltage.

This conclusion is supported by the measured performance of a monolithic IMPATT oscillator [6] incorporating no varactor circuit and coupled line. The IMPATT diode size was equal to that used in the VCO. The monolithic IMPATT gave output power much higher than the VCO. Under the same test condition, 500 mW peak power at 6 percent duty cycle was obtained. Conversion efficiency was 11 percent. All the measured results are referred to the output port of the circuit chip. The insertion loss of the finline transition is accounted for. The much higher output power from the monolithic IMPATT suggests that the VCO circuit has a much lower circuit efficiency. It is hypothesized that the extra loss comes from the varactor circuitry.

## V. Analysis

In order to better understand the VCO circuit, characterized circuit parameters were used in the computer analysis. The capacitance of the varactor diode was measured as a function of voltage. The doping profile of the device was determined by the capacitance-voltage method. The zero bias series resistance was calculated approximately based on the doping profile. The p drift region is 0.3 $\mu$m and is doped to $3.5 \times 10^{16}$. The n region is 0.35 $\mu$m and is doped to $2.5 \times 10^{16}$. The resistance from these two regions is about 14.5 $\Omega$. The metal contact resistance is 1.5 $\Omega$.

Measured capacitance values ($C(V)$) and a voltage-dependent series resistance $R_s(V)$ were used as input for computer analysis. The relationship between $R_s$ and voltage is assumed to be

$$R_s(V) = R_{s0} \times \left(5 - \frac{C_0}{C(V)}\right) \Big/ 4.$$

where $R_{s0}$ is the zero bias series resistance; its value is 16 $\Omega$. $C_0$ is the zero bias capacitance. The factor of 5 is the ratio of the total device length to the avalanche zone length, as measured by $C$–$V$ profiling. This equation assumes that the depletion region extends equally into p and n drift regions at all the bias voltage. Since $C(V)$ is measured, $R_s(V)$ can be calculated accordingly.

These data are inputted into a SUPERCOMPACT program. The coupled line model in SUPERCOMPACT is also modified to give a better fit to the characterized result [5]. The program calculates the resonance frequency, the circuit efficiency, and the load conductance presented to the IMPATT device.

### A. Operation Frequency

Fig. 7 shows the measured oscillation frequency of a VCO for two bias current levels of one IMPATT diode, together with a computer-calculated resonance frequency, as a function of the varactor bias voltage. The agreement between the computer analysis and the experiment is excellent. Below 4 V bias on the varactor, no oscillation can be observed from the VCO. However, the resonance frequency can still be predicted by the computer.

Below 4 V bias voltage on the varactor, the resonance frequency as predicted by the computer becomes fixed. It is caused by the large series resistance of the varactor. Fig. 8 illustrates the effect of series resistance in the model calculation. One curve is the simulation of the experiment, as in Figure 7. The other is the prediction of the model under the assumption of one tenth the series resistance. This value of series resistance would be typical of a single-drift device where only n-type GaAs is used. The tuning range for the small resistance case is 2.87 GHz.

Also shown in Fig. 8 is the measured varactor capacitance at room temperature. The capacitance ratio is only 3, smaller than the maximum available ratio of 5 at high temperature. The undepleted drift region in the varactor is the major source of series resistance.

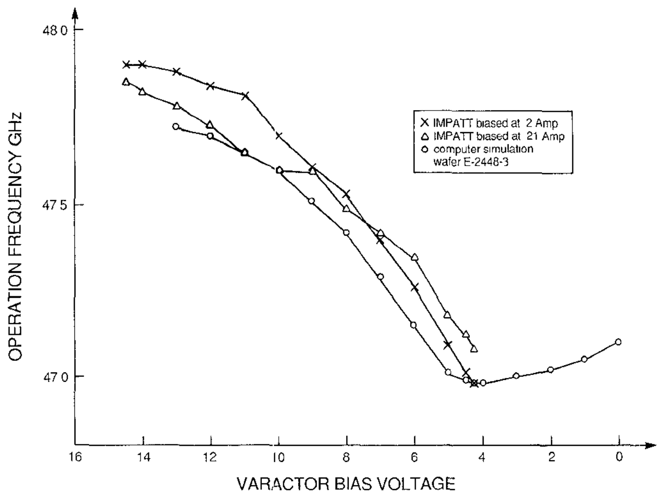



Fig. 7. Operation frequency of a VCO at two bias current levels, together with the result of a computer simulation.

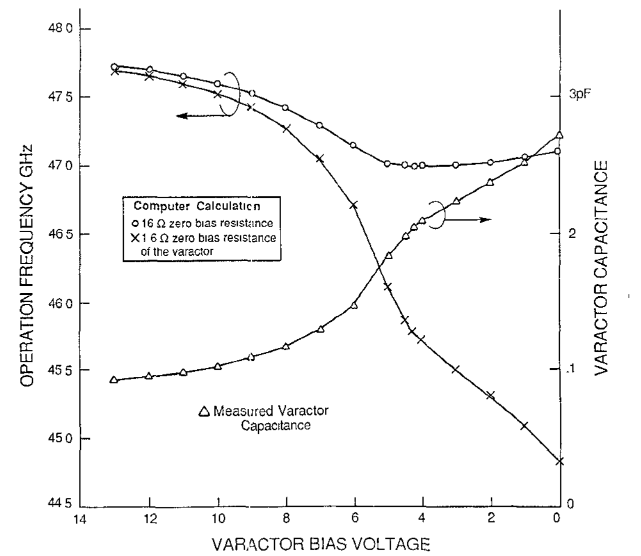



Fig. 8. Calculated resonance frequency with different series resistance values and the measured varactor capacitance as a function of bias voltage.

### B. Load Conductance and Circuit Efficiency

Plotted in Fig. 9 is the load conductance for the two values of series resistance. The low varactor resistance case shows the much lower conductance which would result in oscillator output throughout the entire varactor bias range.

Another important parameter is the circuit efficiency. It is defined as the ratio of the circuit output power to the device-generated power. For a lossless resonator, this number is 100 percent. The circuit efficiency of the VCO was calculated. It is only about 20 percent at 13 V varactor bias. At lower varactor bias voltage (larger series resistance), the circuit efficiency is even lower. The measured circuit efficiency [5] for the monolithic IMPATT chip is much higher, 80 percent. The reduction of the circuit efficiency is caused by both the varactor series resistance and the coupled line structure.

The ratio of the output power of the VCO to that of the

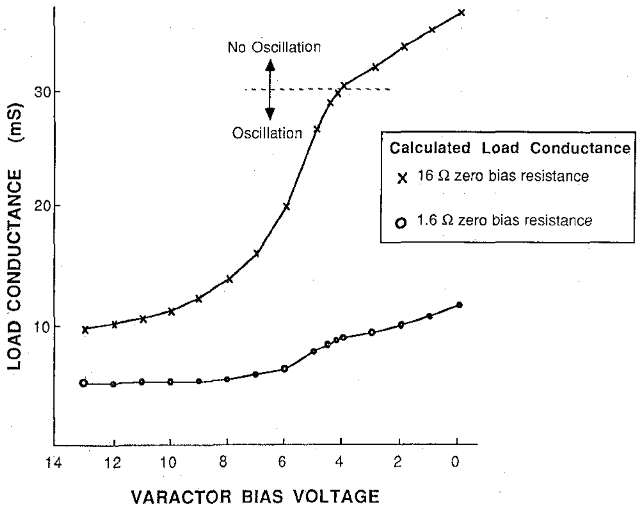



Fig. 9. Calculated load conductance at two different series resistance values.

monolithic oscillator is about the same as the ratio of the calculated VCO circuit efficiency to the measured oscillator efficiency. The measurements imply that the IMPATT diodes in both circuits are well matched. The load impedance versus frequency curve is smooth and has no parasitic resonance loop. A loop in the load impedance curve often causes hysteresis in frequency tuning or the failure of oscillation.

Therefore, the large series resistance of the varactor limits the VCO performance in two ways:

1) It reduces the resonance frequency tuning range as shown in Fig. 8.
2) At low varactor bias voltage, the oscillator is so heavily loaded that no oscillation can take place.

A reduction of the varactor series resistance will greatly enhance the VCO performance.

## VI. Conclusion

A monolithic $Q$-band GaAs IMPATT diode voltage-controlled oscillator has been built and tested. A frequency tuning range from 47 to 48 GHz was achieved with 120 mW peak power. A monolithic IMPATT omitting the varactor circuitry generated 500 mW peak power at 11 percent efficiency. Computer predictions based on measured circuit parameters accurately predict the operation frequency. It is concluded that the key limiting factor in the power and tuning range is the presence of a large parasitic series resistance in the varactor. By utilizing a single-drift doping profile, improved performance can be expected.

## Acknowledgment

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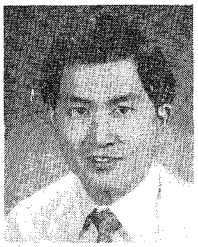

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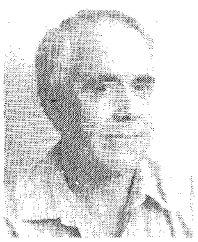

**William Stacey** received his education in England, at Leeds University, where he obtained the B.Sc. degree physics (honors).

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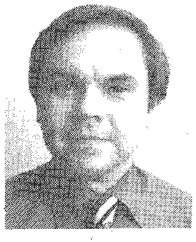

**Ronald C. Brooks** was born in Lynn, MA, on February 21, 1948. He graduated from Wentworth Institute, Boston, MA, in 1970. He received the B.S. degree in industrial engineering from Lowell University, Lowell, MA, in 1984.

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**William E. Hoke** received B.S. degrees in chemistry and physics (with highest honors) from the Pennsylvania State University in 1973. From the University of Illinois, he received the M.S. degree in 1975 and the Ph.D. degree in physical chemistry in 1978.

In 1978 he joined the Raytheon Research Division. The direction of his work has been in exploratory materials research. From 1978 to 1980, he initiated and developed a liquid phase epitaxy effort for growth of GaAs and GaAlAs films. In 1981, Dr. Hoke initiated a metalorganic chemical vapor deposition (MOCVD) effort for the epitaxial growth of CdTe and HgCdTe films. He has also been involved in developing a low-temperature MOCVD HgCdTe technique using new organotellurium sources. In 1981, Dr. Hoke initiated the molecular beam epitaxy (MBE) effort at the Research Division, and in 1988 he initiated a gas source MBE effort for growth of III–V films.

Dr. Hoke has coauthored 20 technical publications, one patent, and five patent applications.