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Ground Surface Sensing Through Plant Foliage

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ABSTRACT

Review of literature indicated that some form of microwave or radar principles were the mostly likely technologies to locate the soil surface through plant foliage. The ground penetrating radar was tested. Both a 120 MHz and 500 MHz system were investigated with the 500MHz giving the best results. This system was able to penetrate the foliage, but the long period between pulses would hinder its use as an input to a control system.

In Israel a radar altimeter system was modified for testing. This system operates in a 2.2 GHz band. At this frequency the system was not able to distinguish between plant material and the surface. Tests using this principle, but at a lower frequencies could not be made because the manufacturer would not release technical information to the investigators.

A third system of a dual frequency continuous wave system was developed specifically for this project. The primary frequency is 1.1 GHz. The system performance did not meet all of the design objectives. The effect of plant material on the reflected signal was more than expected. By using a calibration table for each soil condition, reasonable results could be achieved. Accurate calibration tables would be difficult to devise and their use in the field would require a judgement call by the operator.

Three radar technologies were investigated for sensing the ground surface through plant foliage. None of the three were completely successful. Preliminary data for the reflected characteristics for several soil types and plant conditions were obtained.

FINAL ANNUAL REPORT FOR BARD PROJECT US-1577-88

by

R. B. Dodd, Y. J. Han, U. Pieper and I. Smilovich

GENERAL BACKGROUND INFORMATION

Precise position control of agricultural implements with respect to the ground surface is a very important factor for efficient and effective agricultural production. The wide range of equipment which needs precise height or depth control relative to the ground surface can be classified into several categories.

First, the precise operating depth control in tillage, harvesting and digging machine such as subsoiler, plows and potato, sugarbeet and peanut diggers will save significant amount of energy, will decrease damage to the harvested product, and field losses. Second, onion digger, tomato and cucumber harvesters, sweep cultivators and chemicals incorporation tools are expected to cut roots very close to the ground surface. Third, there is a group of harvesting machines which work above the ground surface. Mowers, grain combines, vegetable harvesting heads, sugarcane harvester, etc., can be classified into this category. For example, when harvesting sugarcane, too low cut introduces dirt and trash into harvester and degrades the product quality. In addition, it causes big soil resistance to the implement. Conversely, if the cut is made too high, some of the product will be left in the field. For all machines which requires precise height or depth control, the ground surface must be measured as close as possible

to the operation point of tools. In modern farming with more and more automatic systems for better quality and uniform productions, the requirement for accurate and automatic height and depth control is expected to be even more critical. Sometimes chemical application machines such as boom sprayers require accurate height control either from the ground surface or from the top of the plant's canopy for better application efficiencies.

Since most of conventional height or depth control was performed by the mechanical and/or hydraulic devices, these systems fall short of optimum operation because of the following reasons: 1) physical dimensions do not allow the sensor to be located close to the actual point of operation, 2) the foliage of the plant and the fruit prevents such a sensor from measuring the ground surface directly, and 3) the interaction and penetration of the mechanical sensors such as direct gage wheel into the soil is not constant along the bed due to varying soil conditions, which causes errors in the measurement for the depth control. To overcome the limitations of the existing sensors, therefore, the adoption of remote sensing concept is demanded.

Some efforts were made to use ultrasonic device for the better precise position control. Paulson and Streliaoff(1974) showed the feasibility of using ultrasonics for depth control of agricultural tools in the field without any obstacle between a sensor and soil surface. Coad et al.(1979) used ultrasonic sensor for the height control of sugarcane harvester's head. With the aid of microprocessor they could enhance the system performance by discriminating of the echo signals from the plant and the soil surface. Nine and Stanley (1980) used air sonar to control the cutter head height of tomato harvester. They concluded only coarse measurements were possible. Because the ultrasonic sensor was not able to detect real ground surface when some obstacles such as vegetation between the sensor and the ground surface existed.

Contrast to the direct mechanical/hydraulic sensing or even acoustic sensing, new technique using microwave was tried. The microwave's penetrating ability through a dielectric medium makes it possible to sense real target's surface under the vegetation obstacles. The advantages of microwave are its directivity and ability of resolving two objects with different reflectivities. Also it is hardly effected by environmental condition such as ambient

temperature.

Numerous researches to identify crop types have been made at various frequency range of microwave. The characteristic of reflectivity of each crop was mainly used to identify crop type and growth. (Paris, 1986; Wu et al.^a, 1985; Ulaby et al., 1983; Zoughi et al., 1986; and Wu et al.^b, 1985). Also, Wu et al. (1975) studied about the microwave reflection from soil surface and various crop types in terms of moisture content. In most of their works the high frequency microwaves in the range of 10 - 20 GHz were used because it was more desirable than the lower frequency microwave in differentiating crop types. Especially Ulaby et al. (1975) reported that even though the sensitivity of the radar backscatter to soil moisture is related to vegetation characteristics and sensor parameters, soil moisture could be sensed through vegetation. They added the lower frequencies provided more information on soil moisture content under vegetation cover due to their inherently better penetrating ability. It is, therefore, induced that low frequency microwave could be a good tool for detecting ground surface even under vegetation.

For those persons interested in this area, a complete listing of publications found are given in Appendix A. This list includes citation in the final report and many other publication which are relevant to the research.

The specific objectives of the project were formulated as follows:

1. Evaluate the potential of using different techniques for remote sensing of the ground surface through plant foliage.
2. Investigate and define the specific parameters and characteristics of microwave reflection from plant and soil material under different agricultural conditions.
3. Formulate and examine a remote sensor system which can identify the surface of the ground through plant foliage for harvesting and/or other agricultural activities. Angle and polarization are classified into the radar parameters. Geometric properties such as density, surface roughness and dielectric property of target are target parameters.

RESULTS AND DISCUSSION

To avoid duplication of effort and to unitize existing hardware, it was decided that the US investigators would use the ground penetrating radar technology and the Israeli investigators would look at other radar technologies. Since all three of the objectives must be met for the system to operate successful, each technology is treated as a seperate project.

GROUND PENETRATING RADAR

System Description

Subsurface Interface Radar(SIR) System-3, Geophysical Survey Systems, Inc. was used through the feasibility test. SIR-3 system is portable broad band pulse radar system, which is able to detect and graphically record the position and depth of objects buried in dielectric materials.

The SIR-3 system consists of transducer, control unit with a built-in graphic recorder and power supply as shown in Figure 1. The signal generated by the transmitter is transmitted into the air through the antenna and the reflected signal from the target surface is received by the antenna. The duplexer enables a single antenna to do the transmitting and the receiving functions alternatively. Then the signal converted into the video frequency in the receiver is fed into the control unit. Passing through the time gain amplifier (TGA), the low pass filter(LPF) and the high pass filter(HPF), the signal is processed for the proper display. In this project, the ground penetrating radar operated only on RADAR mode, so that the graphic recorder was not used.

Transducer

As the most important part of the ground penetrating radar system, the transducer consists of transmitter, receiver and antenna. TR-3102 Transducer, which has a center frequency of 500 MHz, was used through the test. The transmitter radiates the high-power and short pulse signal through the antenna. Usually a group of pulses are transmitted periodically to increase the output signal-to-noise ratio and to enhance target detection. This pulse repetition frequency (PR) is determined primarily by the maximum range where targets are expected, because the time interval between pulse transmissions should be long

Figure 1. SIR-3 ground penetration radar and its functional block diagram.

enough not to transmit a signal before the echo signal from the target returns.

When the antenna of ground penetrating radar was placed approximately 137 cm above the soil surface, it takes 9 nanoseconds for the return signal to come back from the target surface. The PR of 51.2 KHz in the ground penetrating radar setup provides sufficient time for the surface return signal. The maximum transmitting power reaches 2.5 kW at this pulse repetition frequency.

The receiver converts the returning pulse, which is only several nanoseconds in duration, into analog signal of tens of milliseconds in duration, and returns this analog signal to the ground penetrating radar control unit. The signal was then processed and sent to the display device. Since the scan rate of 16 scans/sec and the PR of 51.2 KHz were used through the test, the actual signals picked up by the antenna are repeated 3200 times across a single scan. The periodicity of the received signals allows a single sample to be taken from pulse-to-pulse to reconstruct the real-time signal over a longer time period. This is done by taking samples from the real-time waveforms at times differing from the transmit pulse. The 400 clock pulses during a single scan generated by the timing clock of 6.4 KHz in the ground penetrating radar are used to provide time-delay to the sampler. Each clock pulse delays the next sampling strobe on the real-time waveform from the previous sampling strobe. Therefore, a single sampling occurs at 400 different positions across one real-time waveform. At each of these 400 positions, 8 samples are taken and integrated to increase the output signal-to-noise ratio. The time delay of sampling strobe at each clock pulse depends on the range setup in the ground penetrating radar. Although the scan rate of the system was set to 16 scans/sec, which gives 62.5 millisecond in reconstructed signal, the corresponding time-period in real-time waveform can be adjusted by the range setup in the ground penetrating radar. When the target surface is detected at the time T_c in the constructed waveform, the actual elapsed time for the microwave's round-trip to target, T_w in nanosecond becomes

$$T_w = \frac{S_r \cdot T_c}{62.5} \quad (1)$$

where, S_r is the range setup of the ground penetrating radar (unit : nanoseconds).

Thus, the distance from the antenna to target surface, R in cm can be obtained as

$$R = c \cdot \frac{T_w}{2} \quad (2)$$

where, c is the microwave propagation speed in the air (30 cm/nanosecond).

The antenna has a fairly broad radiation pattern, which is conical in shape with the apex of the cone at the center of the antenna. The beam width of antenna is approximately 90° inclusive angle in the fore-aft direction and 45° in the sides direction.

Control Unit with Built-in Graphic Recorder

The main function of control unit is not only to supply power and synchronizing signals to the transducer, but also to set up the operating parameters for the transducer. The return signals coming from the receiver are also processed by the control unit for the display.

The control unit has two built-in time gain amplifiers, which make it possible to adjust gain separately over the near-, the intermediate- and the far-range in a single scan. In addition, a low pass filter and a high pass filter are employed for the signal processing. After gain adjustment and filtering, the microwave signal is sent through I/O connector of the control unit to external device such as oscilloscope or microcomputer.

Detection of Soil Surface

Even though the ground penetrating radar has been originally made for the subsurface survey, information of the altitude of antenna above soil surface can be obtained.

The distance to a reflecting target such as soil surface is indicated by the elapsed time between the transmission of a microwave signal and the reception of the echo from soil surface. Then the elapsed time can be converted into the distance because the propagation speed of the microwave is known.

The output signal of ground penetrating radar consists of transmitted pulse, surface return and returns from the interfaces below soil surface. Since the transmitted pulse always remains constant as long as the range setup of radar

system are not changed, it can be served as a time reference. Therefore, the elapsed time between the transmitted pulse and the surface return can be obtained.

The output signal come from the ground penetrating radar is a synthetic radar signal, which is the convolution of the actual reflected signal with the unit sample response of the system. Figure 2 shows a typical ground penetrating radar output from a digital oscilloscope (NIC 310, Nicolet Co.). The output was obtained for a single scan of 62.5 milliseconds. The reflected signal was composed of a group of high-amplitude signals, which is caused by oscillations within the antenna due to the characteristic of a particular antenna. Since microwave is always reflected at the interface between two different media, the reflected signal from soil surface must be the first big one after the transmitted pulse.

For example, suppose that the reflection from soil surface occurs at 120th sample in the discrete waveform of the signal picked up by the receiver during a single scan as shown in Figure 3a. The unit sample response of the system could be assumed as shown in Figure 3b. The convolution of these two signals are defined as follows:

$$x(n) = x(t) \text{ at } t = n \cdot T \quad (3)$$

$$y(n) = x(n) * h(n) \quad (4)$$

where, n : the sample number in integer, $y(n)$: the synthetic radar signal, $h(n)$: the unit sample response of the system, $x(n)$: the real reflected signal and T : the sampling period.

In Figure 3c, the result of convolution shows that the difference in sample numbers between the first peak of the transmitted signal group and the first one of the surface return is the same as the difference between the transmitted pulse and the reflection from the soil surface in Figure 3a.

Therefore, the elapsed time between the transmitted pulse and the surface return can be determined by measuring the time difference between the first positive peaks of the transmitted pulse and the surface return, respectively.

Another characteristic of the ground penetrating radar enhancing its performance is that the two-way reflection sensitivity of the antenna is small for flat surface at normal incidence of microwave beam. Even though the antenna has a fairly broad radiation pattern, the microwave reflected at any surface not

Figure 2. Typical output signal of the ground penetrating radar.

Figure 3. Synthetic radar signal: (a) assumed reflected signal, (b) assumed unit sample response of the system, and (c) convolution result.

normal to the incident microwave beam is not received. This characteristic enables the ground penetrating radar system to measure the distance to soil surface accurately.

Reflection of Microwave

Microwave reflection is largely dependent on the electrical characteristic such as the conductivity of target material. To be detected by the radar receiver, the reflecting medium should have sufficient conductivity so that the amount of energy of reflected signal is greater than the system noise.

The effect of target materials on the microwave reflection was investigated experimentally on different media such as galvanized sheet metal, wet soil surface and dry concrete floor. The experiments were conducted at the same combination of gain amplifiers in the ground penetrating radar system and the same distance to the target surface. As shown in Figure 4, the amplitude of surface return signal was affected by the conductivity of the reflected media. The galvanized sheet metal was a complete conducting material and the dry concrete floor surface was very close to the complete dielectric material. Since the soil moisture content was relatively high in the experiment, the wet soil surface gave stronger return on its surface than on the dry concrete surface.

Theoretically, the reflection coefficients of microwave on the above three target materials can be obtained from equation 2 by using the impedances of target materials from the last column in the Table 1. The impedance value Z 's were calculated from equation 3 with the electrical parameters of some natural materials presented in the Table. Compared with the reflection coefficient of almost 1 for the iron surface, the reflection coefficients on the concrete floor and wet soil surface were 0.42 and 0.63, respectively. These calculation results agreed with the experimental results shown in Figure 4.

Effective Illuminated Area

The effective illuminated area on the target surface and the effect of target characteristics on the microwave reflection were investigated by measuring the amplitude of surface return signal on the highly conductive surface target such as galvanized sheet metal over the less conductive surface such as concrete floor. The ground penetrating antenna was located 137 cm above concrete floor.

Figure 4. Comparison of the amplitudes of reflected signal on (a) galvanized sheet metal, (b) wet soil surface, and (c) dry concrete floor.

One side of galvanized sheet metal of 66 cm x 89 cm was placed and aligned with the center line of the imaginary illuminated area on the concrete floor under the ground penetrating radar antenna. As the galvanized sheet metal was moved outward, the amplitudes of surface return signals were measured by a digital oscilloscope.

The amplitudes of surface return signal were plotted in term of the antenna half-beam width. As shown in Figure 5, the amplitude of the return signal was reduced as the galvanized sheet metal was moved outward from the center line. After the antenna half-beam width exceeded 15°, the amplitude remained constant, which indicated that the target was out of the effective illuminated area. Therefore, the effective illuminated area, A_e , can be expressed in terms of the distance from the GPR antenna to soil surface.

$$A_e = \pi \cdot [h \cdot \tan(15^\circ)]^2 \quad (5)$$

where, h is the distance from the antenna to soil surface.

Effect of Edges on Terraced Surface

In order to investigate the radar reflection at the edge area on the terraced surface, the galvanized sheet metal of 66 cm x 89 cm was placed 8.5 cm above the concrete floor and at the center line of the illuminated area in the laboratory. Moving the galvanized sheet metal outward from the center line, the surface return signals were measured by the digital oscilloscope. The distances from the GPR antenna to target surface were plotted in terms of the galvanized sheet metal locations, which were expressed as the antenna half-beam width. As shown in Figure 6, while the galvanized sheet metal was located within approximately 15° of antenna half-beam width, the ground penetrating radar output indicated a median value between the distance to galvanized sheet metal surface and the distance to concrete floor. Once the galvanized sheet metal was completely moved out of the effective antenna beam width, the ground penetrating radar output showed only the distance to concrete floor. This indicates that the distance to target surface read by the GPR on the terraced surface is a midpoint between the maximum and the minimum distance within the effective illuminated area.

Figure 5. Effective antenna half-beam width of the ground penetrating radar.

Figure 6. Effect of edges of terraced surface on the surface return.

Effect of Moisture Content

Since the moisture content of a medium is directly related to the conductivity, the moisture content of soil surface is one of the most important factors to affect the reflection of radar signal. In order to investigate the effect of soil moisture content on the microwave reflection, the radar responses were taken at several different soil moisture content. For the experiment, dry soil with high bulk density was packed into a 1.22 m x 1.22 m x 0.075 m wooden box. The initial gravimetric moisture content of soil was 7.47 % dry basis (db) and the bulk density was 0.86 g/cm³. After one liter of pure water was evenly sprayed into the soil box, the radar response was taken again. This process was repeated up to 13 liters of added water. The final gravimetric moisture content was 26.74 % (db).

As shown in Figure 7, in the range of lower moisture content, the amplitude of reflected signal tends to increase as the moisture content of soil increases. It is a natural tendency of the radar signal because the conductivity of soil increases as the moisture content. Beyond the moisture content of 12 % (db), there was no significant change in the amplitude of signal.

Effect of Vegetation Material

The effect of vegetation on the microwave reflection was investigated experimentally in the laboratory and in the field.

Laboratory Test

Two bales of hay were placed on the concrete floor in the laboratory under the ground penetrating radar antenna and the reflected signals were collected through the digital oscilloscope. The physical properties of hay materials are presented in Table 2.

Figure 7. Effect of the moisture content on the reflected signal strength.

Table 2. Physical properties of hay materials.

	A	B	C
Material	grass	grass	barley straw
Size(cm)	53x38x20	53x38x40	91x43x45
Density(g/cm ³)	0.04	0.04	0.07

No effect of vegetation on the target surface return was observed in the hay materials A and B. However, the hay material C had an effect on the surface return signal. As shown in Figure 8, the surface return was delayed with respect to the surface return without vegetation, which makes the measured distance to target longer than the actual. This was due to the fact that the vegetation slowed down the speed of microwave propagation. The propagation speed of microwave inside dielectric medium, v_m , becomes:

$$v_m = \frac{c}{\sqrt{\epsilon_r}} \quad (6)$$

where, c is the propagation speed of microwave in air and ϵ_r is the dielectric constant of the medium.

This time delay results in an error in measuring the distance from the antenna to the target surface. In the hay material C, the amount of delayed time was 0.45 millisecond in the reconstructed waveform, which corresponded to the error of 2.2 cm in the distance measurement.

The experimental results indicated that the vegetation material did not affect the system performance except the case of very heavy vegetation.

Field Test

During June 1991, some experiments were conducted in the various target characteristics such as agricultural field with tall plant such as foxtail, cultivated soil and road paved with small gravel. The antenna was located 137 cm above those target surfaces, the ground penetrating radar indicated exact distance from the antenna to target surface in above all cases. Regardless of plant and/or vegetation, the distance to soil surface were able to be measured accurately.

Figure 8. Effect of heavy vegetation hay material C on the surface return.

The experimental results showed that the GPR with the operating frequency of 500 MHz can be used to measure the distance to the soil surface through vegetation.

Real-Time Data Acquisition

The SIR-3 ground penetrating radar system was interfaced with a real time data acquisition system to be used as a remote sensor for the height/depth control of agricultural implements. Lab Master DMA (Scientific Solutions, Inc.) board was installed on a microcomputer for data acquisition. The Lab Master board consists of mother board and daughter board, which can perform A/D conversion up to 100 KHz through DMA operation.

The ground penetrating radar system outputs a pre-processed radar signal with a sweep signal. The sweep signal was connected to one of the counters on the board to trigger the start of A/D conversion. The radar signal was connected to the one of the analog input channels on the Lab Master board.

Detection Algorithm

The received analog signal was converted to a digital signal during a single scan and was stored in the memory buffer through DMA operation. After completion of A/D conversion, the data in the buffer were processed to find out the peak of surface return signal.

The sampling frequency of A/D conversion was set to 20 KHz and an array of 1×1250 was generated in the memory buffer. The A/D conversion of the radar signal started at the falling edge of the sweep signal and stopped at the end of 1250 A/D conversions. While the peak detection procedure was proceeded, the counter kept watching the sweep signal. The flow chart of the algorithm for the data acquisition and processing is presented in Figure 9.

These procedures were repeated on every fourth sweep signal. Data acquisition was completed during the first single scan, which is 62.5 msec long, and the calculation to determine the distance from the antenna to target surface was proceeded during the following two single scans of 125 msec. Therefore, the actual data was sampled at every 187.5 msec.

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Figure 9. Flow chart of data acquisition and processing for soil surface detection with the ground penetrating radar.

The distance to soil surface was determined at the first positive peak in a group of relatively high-amplitude signals after the transmitted signal. Since the transmitted signal always appears at the same position in the time-domain, the search procedure starts after a fixed time to skip the transmitted pulse. The peak detection was performed on the basis of a predetermined threshold value.

The peak detection would be simple if the amplitude of radar signal is over the threshold value and increases monotonically up to the first peak. In practice, however, some local maxima were observed because of the signal fluctuations. To avoid such a local maximum value being picked up as a true peak, the nonequal weighted running average technique was used as a low pass filter. This technique is similar to the simple moving average except giving more weight to the most recent sample, which could be described by the following equation (Coad et al., 1979):

$$Av(n) = Av(n-1) + \frac{S(n) - Av(n-1)}{M} \quad (7)$$

where: M is the weight index,

$S(n)$ is the sample value of n^{th} sample, and

$Av(n)$ is the accumulated average n samples have been taken.

The value of M is adjustable for the different surfaces so that the peak detection algorithm detects general changes in soil surface profile but not to be too sensitive to the short irregularities on soil surface and even to small measurement error of the radar system. The positive peak is determined at the sample number where the $Av(n)$ is no longer increasing in value. The elapsed time T in milliseconds between the transmitted pulse and the surface return signal is determined by

$$T = n \cdot \frac{1}{f_s} - 6.50 \quad (8)$$

where: n is the sample number and

f_s is the sampling frequency.

Lab Experiment

For preliminary evaluation of the system's ability in the real time operation, a test was conducted over a prepared soil bin located in Soil Dynamics Laboratory at Clemson University. A moving table of 3 m x 1 m x 0.15 m of the soil bin travels in a linear motion on the track of 12 m in length. An artificial edge profile was made in the middle of prepared soil surface. The distance between two transition was 0.6 m. The height difference between two soil surfaces was approximately 5 cm. The GPR antenna was hung on the fixed cantilever above the soil surface. The moving soil bin under the ground penetrating radar antenna simulated the antenna passing over a soil surface. The range and the scan rate of the SIR-3 ground penetrating radar were set to 20 nanoseconds and 16 scans/sec, respectively. The velocity of the cart was set to 0.143 m/sec. Therefore, the data were taken on every 2.68 cm along the soil surface profile. The weight index of peak detection algorithm was set to 4.

Test result was shown in Figure 10. Although the ground penetrating radar system showed the overall profile of soil surface, it could not detect the sharp transition of the soil profile.

Field Experiment

The antenna was suspended by a cantilever 1.5 m away from the truck bed to avoid a possible reflection from the truck. Field experiment have been conducted on the barley field in Simpson Experimental Station, Pendleton, SC. A piece of agricultural field was divided by several small plots and barley was being grown in every other plot. Bare soil surface exists between barley plots. The average height of barley was about 84 cm. The experimental site and setup are shown in Figure 11.

The radar output signals were recorded by the digital oscilloscope at 5 spots each on the bare soil surface and on the soil surface covered with barley. The distance to soil surface was determined from the surface return signals using the developed detection algorithm. The actual distances between the ground penetrating antenna and the soil surface on each spot were measured by a tape measure.

The experimental results were summarized in Table 3. The measurement error ranged from 0.48 cm to 6.5 cm, which might be caused by the measurement error by

Figure 10. Experimental result of real-time data acquisition and processing by the detection algorithm.

Figure 11. Experimental site and set up for the field test.

Table 3.3. Measurement error between the radar system and the tape measure on bare soil surface and on crop field.

	True distance	Measured distance	Difference	Error(%)
Soil 1	180.98	178.20	2.78	1.54
Soil 2	181.61	177.22	4.39	2.42
Soil 3	182.88	176.35	6.53	3.57
Soil 4	174.63	176.83	-2.21	1.27
Soil 5	179.07	174.98	4.09	2.28
Crop 1	176.53	179.76	-3.23	1.83
Crop 2	179.07	178.59	0.48	0.29
Crop 3	176.53	178.20	-1.67	0.95
Crop 4	175.90	177.32	-1.42	0.81
Crop 5	174.63	177.22	-2.59	1.48

Unit : cm

the tape measure because of soil surface roughness. It was observed that the power amplitudes of surface return signals were consistently lower on the soil surface covered with barley than on the bare soil surface, which indicated that there were attenuation due to barley. Although the signal strength was attenuated by the crop, it was strong enough to be detected without affecting the distance measurement. The t-test result showed that there was no significant difference at 95% confidence level between the actual distance and the distances measured by the GPR.

ISRAELI DEVELOPED RADAR

Due to a previous research between the Technion and Elta, Elta was contacted for assistance with the project. Elta is a defense contractor which has a high level of expertise in radar technology. However due to security reason, they were very reluctant to become involved. After several lengthy discussions, they agreed to supply a modified unit for limited testing. All details on how the system operated and the actual frequencies involved were consisted to be confidential by Elta and were not available for discussion. All of the work accomplished on this system treats the radar as a "black box" with an output.

The physical description of the system is very limited because of the uncooperative nature of the manufacturer. The system is very compact and rugged. The cost of the system would be totally out of range for an agricultural application. The cost of the system as tested was rumored to be \$100,000 dollars. However, since this project was a proof of concept, we decided to go ahead with testing the system. If the system was successful, it was rationalized that the major cost for the system was meeting the stringent military requirements and a cheaper unit could be produced.

The Israeli system utilizes only one antenna approximately 0.3 m in diameter. The operating frequency is in the range of 2 GHz. The hardware signal analyzing device produced a digital output to the computer. The design allowed two working modes for static and dynamic measurements.

The static mode gives the signal in degrees of intensity as a function of the height of the antenna from the surface. A computer program collects 10 readings in the static test and records the average signal. In the dynamic mode, only a digital output correlating to the height is recorded at a rate of 20 samples per seconds. A special fixture was made to hold the antenna at different heights and angles relative to the ground level. A different fixture was fabricated to mount the antenna on a tractor for dynamics testing.

Preliminary tests were conducted with the Elta radar system. Static tests checked the influence of soil conditions. Three type of soils were tested, two

clays and a sandy soil. Each has two degrees of water content. In one clay soil, four different clod sizes were sampled. In all tests a better signal and more accurate sensing of height readings were noted for wet soil conditions. All soil surfaces indicted a similar signature. A variation in the reflected signals measuring the height in the same soil type, but from various clod sizes, raised the question as to what is the real surface level. Tests in salty soil are needed.

In the second stage, static height measurements of several plants were conducted. The test site was located in the Botanic-ecology garden at the Technion. Only in the cases of dry type plant foliage was an accurate measurement of the antenna height achieved. The plant configuration as well as its moisture content had a very high influence on the results. The third stage checked the dynamic mode of the system. The radar and all its hardware was mounted on a tractor. A contact linear transducer was connected to record the actual heightt of the antenna. The readings from the radar and linear contact transducer were collected at a rate of 15 samples per second. When tested on a hard surface the system gave very good results. A series of boxes with varied heights were arranged in a line in a soil bin at the Technion. With dry sandy soil, the system was able to draw a good approximation of the profile of the boxes. When fleshy plant material was placed on the top of the boxes, an accurate measurement of the soil surface could not be made.

Due to the desires of Elta to keep information about the system from being made public only a limited general description of the system would be released. Whether this is due to industrial or military reasons is unclear to the investigators. The system has potential , but other frequencies must be investigated. Without the support of a company with expertise in this area, the chances of success are greatly lessened.

DESIGN OF DUAL-FREQUENCY CONTINUOUS-WAVE RADAR

Principles of Operation

When a signal of sinusoidal form $\sin 2\pi f_0 t$ is radiated from a radar system, it travels to the target at a distance R and returns to the radar system after a time, τ .

$$\tau = \frac{2 \cdot R}{c} \quad (9)$$

where, c is the microwave propagation speed in the air.

Since the signal reflected on the target surface and received by the radar system has been delayed for time τ , it can be expressed by $\sin[2\pi f_0(t-\tau)]$. There exists the phase difference of $2\pi f_0 \tau$ with reference to the transmitted signal. The phase difference ϕ is

$$\phi = 4\pi \cdot \frac{f_0 \cdot R}{c} \quad (10)$$

$$= 4\pi \cdot \frac{R}{\lambda} \quad (11)$$

where, f_0 is the operating frequency and

λ is the wavelength ($\lambda = c/f_0$).

This phase difference is proportional to the distance to target R . Therefore, the distance to target can be measured by measuring the phase difference between the transmitted and the received signal. However, the value of ϕ in the above equation can not be greater than 2π radian in order to avoid the phase difference being ambiguous. This means that the maximum range to be measured unambiguously becomes only one-half of the wavelength. At microwave frequencies, this unambiguous range is too small to be practical. This is the reason why the ordinary continuous-wave radar is not being used for the range measurement.

The region of maximum unambiguous range can be extended considerably by utilizing two different frequencies of continuous-wave signals. Since the dual frequency continuous-wave radar system transmits two different frequency continuous-waves simultaneously, both continuous-waves are in phase at the zero

range. As two microwaves propagate outward from the radar system, the relative phase difference between two different frequency continuous-waves increases gradually. Therefore, the phase difference of the received signals of frequencies f_1 and f_2 at an instantaneous distance R_i becomes

$$\phi_n = \frac{4\pi \cdot f_1 \cdot R_i}{c} \quad (12a)$$

$$\phi_a = \frac{4\pi \cdot f_2 \cdot R_i}{c} \quad (12b)$$

The instantaneous distance R_i can be obtained by subtracting equation (12b) from equation (12a):

$$\Delta\phi = 4\pi \cdot R_i \cdot \left(\frac{f_1 - f_2}{c} \right) \quad (13)$$

where, $\Delta\phi = \phi_n - \phi_a$.

Therefore,

$$R_i = \frac{c \cdot \Delta\phi}{4\pi \cdot \Delta f} \quad (14)$$

where, $\Delta f = f_1 - f_2$.

Since Δf is constant, the distance to target becomes a function of the phase difference between two received continuous-waves at two different frequencies f_1 and f_2 .

Since the phase difference is periodic, equation (14) is meaningful only within the range of $0 < \Delta\phi < 2\pi$. Therefore, the maximum unambiguous range becomes

$$R_{\max} = \frac{c \cdot 2\pi}{4\pi \cdot \Delta f} = \frac{c}{2 \cdot \Delta f} \quad (15)$$

This range is repeated on every 2π in phase difference. In other words, the maximum unambiguous range is repeated periodically as the microwave propagates. One maximum unambiguous range can be considered as a window in which the distance to target can be measured unambiguously from the base point of window. For example, if a target is located far away from the radar system, the measuring

window can be moved up to the range which includes the target.

The equation (15) represents the theoretical unambiguous range when the phase shift can be detected between 0 and 2π . In practice, however, most phase detectors on the market only measures the phase shift between 0 and π . In this case, the practical unambiguous range becomes one-half of the maximum unambiguous range defined in equation (15). Hence,

$$R_p = \frac{1}{2} R_{\max} = \frac{c}{4 \cdot \Delta f} \quad (16)$$

where, R_p is the practical unambiguous range.

The theoretical rms range error, ϵ_r , in the dual frequency continuous-wave radar is (Skolnik, 1980):

$$\epsilon_r = \frac{c}{4\pi \cdot \Delta f (2E/N_0)^{1/2}} \quad (4.8)$$

where, c is the propagation speed of microwave, E is the energy contained in received signal, and N_0 is the noise power per Hz of band width.

Also the range resolution may be estimated by the following relationship (Miller et al., 1984):

$$\sigma_\phi^2 \geq \text{SNR}^{-1} \quad (18)$$

where, σ_ϕ is the rms phase uncertainty in radian, and SNR is the signal-to-noise ratio.

The value of σ_ϕ translates into a distance of ΔR as the lower bound uncertainty for a single observation.

$$\Delta R = \frac{R_{\max} \cdot \sigma_\phi}{2\pi} \quad (19)$$

Design Parameters of Radar System

Operating Frequency

The desired operating frequency depends on the microwave's penetrability

through plant/vegetation and physical size of the antenna. The preliminary test conducted at Clemson University using a ground-penetrating radar showed that the operating frequency of 500 MHz could sense the true soil surface through vegetation. When the operating frequency is too high, the reflection at the soil surface becomes no longer dominant because of the increased attenuation due to vegetation. On the other hand, lower operating frequency requires a larger antenna because the physical size of the antenna is inversely proportional to the operating frequency. In this project, the operating frequency of 1.1 GHz was used as a compromise between these two requirements.

Frequency Separation

The frequency separation Δf between the two frequencies directly affects the unambiguous range and the measurement error. Equation (17) indicates the measurement error can be reduced by increasing the frequency separation Δf . However, the greater the frequency separation, the shorter the maximum unambiguous range as shown in equation (15). In addition, too wide separation may result in poor impedance matching of the antenna at the modulated frequencies because the antenna has been matched to the system at the carrier frequency of 1.1 GHz. The selection of Δf , therefore, represents a compromise between the unambiguous range, the requirement of measurement accuracy and the characteristics of system.

From equation (16), the required frequency separation becomes

$$\Delta f = \frac{c}{4 \cdot R_p} \quad (20)$$

Assuming that the practical unambiguous range of 30 cm is sufficient to measure the height variation of soil surface, the frequency separation was rounded to 240 MHz in this study. The practical unambiguous range in this case became 31.25 cm.

The theoretical range error from equation (17) became 0.5 cm assuming the signal-to-noise ratio (E/N_0) = 200 (Miller et al., 1984). The lower bound of uncertainty for a single observation from equations (18) and (19) was 0.71 cm. In this case, the dual frequency continuous-wave radar is able to theoretically measure the distance to target with a resolution of 0.71 cm and a range error of 0.5 cm.

System Integration

A block diagram of the proposed radar system is shown in Figure 12. In selecting and integrating the components of the radar system, the special attention was given to the dynamic range of each component as well as the operating frequency range. The mismatching between components causes the power reflection back to the preceding component, which deteriorates the performance of components, and possibly causes a permanent damage to the components.

Generation and Transmission of Microwave Signal

Two oscillators are used as signal sources for the radar system. The first oscillator generates a carrier frequency of 1.1 GHz and the second one, called local oscillator, generates a local frequency of 120 MHz. .

A balanced modulator modulates the carrier signal with the local frequency to generate two identical, but different frequency signals as follows.

$$f_1 = f_o - f_m \quad (21a)$$

$$f_2 = f_o + f_m \quad (21b)$$

where, f_o is the carrier frequency and

f_m is the modulating frequency.

A double-balanced mixer (Mini-Circuit, ZFM-4H) was used as the balanced modulator. Applying the carrier signal w_c to the R port of the double-balanced mixer and the modulating signal w_m to the IF port, the carrier signal is suppressed and its side-lobe is amplified. The modulating signal centered on the suppressed carrier signal is taken from the LO port of the double-balanced mixer.

The high isolation of the double-balanced mixer between ports enables higher suppression of the carrier signal while the undesired higher-order sidebands are almost totally removed.

Assuming the generated signal has a sinusoidal form, the output of the balanced modulator, V_{bm} can be expressed as

Figure 12. Block diagram of dual-frequency continuous-wave radar system.

$$V_{\text{tm}} = \sin(w_1 \cdot t) + \sin(w_2 \cdot t) \quad (22)$$

where, $w_1 = w_c - w_m$ and $w_2 = w_c + w_m$.

The modulated output signals are transmitted through a transmitting antenna.

A separate 240 MHz reference signal was generated by doubling the signal from the 120 MHz local oscillator. The reference signal was used as a reference to detect the phase shift of the received signal.

Reception and Detection of Returning Signal

The microwave signals reflected on the soil surface are picked up by the receiving antenna. The received signal V_r becomes

$$V_r = \sin(w_1 \cdot t + \phi_1(t)) + \sin(w_2 \cdot t + \phi_2(t)) \quad (23)$$

where: $\phi_1(t)$ is the phase difference at the signal of operating frequency f_1 and

$\phi_2(t)$ at the signal of operating frequency f_2 .

The raw signal received by the receiving antenna was too weak to process, possibly due to the propagation loss, attenuation by plant/vegetation and scattering. The raw signal was amplified by a power amplifier (Miteq Co., AM-3A-0515). The amplified signal was fed into a square law detector (MA-Com, Backdiode-Detector). The following current-voltage characteristic was used in the detector (Zeines, 1970).

$$i = a_1 \cdot v + a_2 \cdot v^2 \quad (24)$$

Since only the frequency changes through the system are of primary concern, in the following discussion, the constants a_1 and a_2 are omitted for the convenience.

When V_r is applied to the input terminal of the square law detector, the output current becomes

$$i = \{\sin(w_1 \cdot t + \phi_1(t)) + \sin(w_2 \cdot t + \phi_2(t))\} + \{\sin(w_1 \cdot t + \phi_1(t)) + \sin(w_2 \cdot t + \phi_2(t))\}^2 \quad (25)$$

Defactorizing the above trigonometric expression, the output waveform contains various frequencies as shown in the following table.

Term	Frequency
dc	0
carriers	w_1 & w_2
carrier 2nd harmonic	$2 \cdot w_1$ & $2 \cdot w_2$
original side frequencies	$(w_1 + w_2)$ & $(w_1 - w_2)$

The square law detector filters out the frequency terms w_1 , w_2 and the higher frequencies so that only $(w_1 - w_2)$ frequency term passes through. Therefore, the output signal from the square law detector can be expressed as

$$V_d = \cos[(w_1 - w_2) \cdot t + (\phi_1(t) - \phi_2(t))] \quad (26)$$

This signal is amplified for the further processing. The specification of the amplifier for this purpose is Miteq Co., AU-2A-0150. The amplified signal becomes

$$V_s = \cos[(w_1 - w_2) \cdot t + (\phi_1(t) - \phi_2(t))] \quad (27)$$

The above signal and the reference signal, mentioned earlier, were compared in a phase detector. A double-balanced mixer (Mini-Circuit, ZFM-2) was used as the phase detector. When two signals of identical frequency but different phase are applied to the R and LO ports of the double-balanced mixer, dc voltage proportional to the phase difference will be output at IF port of the double-balanced mixer.

There are some advantages in using the double-balanced mixer for the phase detector. The balanced configuration of the double-balanced mixer reduces input

signal interaction and eliminates the necessity of expensive, time consuming null adjustments. However, they require the two input signals to be approximately same amplitude level and within the amplitude range of 0 to 7 dBm at the same time. The higher input level above the optimum amplitude range will tend to produce an unbalanced characteristic while lower input levels may deliver an insufficient dc output voltage. To obtain the output which is independent of input signal variations, the double-balanced mixer should operate in a saturated mode, which requires sufficiently high input signal levels. Otherwise, the output will vary according to the amplitude variation as well as the phase difference.

This phase detector produces output voltage in the range of -300 mV to 300 mV for the phase difference of 0° to 180° . The detected phase difference $\phi_d(t)$ becomes

$$\phi_d(t) = \phi_r - [\phi_1(t) - \phi_2(t)] \quad (28)$$

where, ϕ_r is a phase of the reference signal.

The value of $\phi_d(t)$ can be adjusted by shifting the phase of reference signal ϕ_r , since the absolute phase difference $[\phi_1(t) - \phi_2(t)]$ is not affected by changing ϕ_r . Therefore, the distance to target surface can be measured by the output voltage of the phase detector.

Antenna

As a mean for radiating and receiving microwaves, antenna takes various forms to meet the particular need. In this project, a parabolic cylinder antenna was adopted because of its simplicity and low cost to manufacture. It consists of a cylindrical reflector with parabolic cross section and a line feed along the focal line of the reflector. The half-wavelength dipole was used as a line feed source for the antenna. The configuration of parabolic cylinder antenna is shown in Figure 13.

Primary Antenna - Dipole

Dipole antenna is known as the one of the simplest, cheapest and the most versatile for many applications. Especially, the half-wavelength dipole antenna

Figure 13. Configuration of a parabolic cylinder antenna.

is commonly used because its theoretical radiation resistance of 73 ohms is close to 50 ohms characteristic impedance of a typical transmission line.

The characteristic of dipole antenna can be described in terms of the antenna parameters such as the radiation pattern, directivity, gain and radiation resistance.

Radiation Pattern

Generally the radiation pattern of the antenna is defined as the relative distribution of radiated power as a function of direction in space. Dipole has an omnidirectional radiation pattern, which has a nondirectional pattern in azimuth (H-plane) and a directional pattern in elevation (E-plane). The radiation pattern characteristic of dipole antenna is, therefore, expressed in terms of only angular displacement on E-plane on both electric and magnetic field. The theoretical half-power beam width of the half-wavelength dipole is 78° on E-plane.

Directivity and Gain

As a measure of the ability of an antenna to concentrate the radiated power in a given direction, the directivity of antenna is defined as the ratio of the power radiated per unit solid angle to the average power radiated per unit solid angle. For the half-wavelength dipole the maximum directivity is 1.64 theoretically.

The gain of an antenna is proportional to its directivity and can be expressed as

$$G = \eta \cdot D \quad (29)$$

where: G is the gain of the antenna,

D is the directivity and

η is the antenna efficiency.

For the perfectly efficient antenna, $\eta=1$ so that the gain is equivalent to its directivity. Thus, the theoretical gain of dipole antenna can not be greater than 2.2 dB.

In practice, however, the antenna efficiency ranges between 50% and 75% which limits the gain of the dipole antenna to less than 0.9 dB.

Radiation Resistance

The effectiveness of an antenna could be expressed by the amount of real power the current I_0 can generate.

$$P_t = 2 \cdot I_0^2 \cdot R_{rad} \quad (30)$$

where: P_t is the transmitted power, and R_{rad} is the radiation resistance.

Since the transmitted power is proportional to the radiation resistance, the radiation resistance should be large enough to radiate the sufficient amount of power.

To be efficient, the dipole should have a comparable length to the wavelength of the microwave, because the radiation resistance in the dipole antenna increases as the length of dipole increases. For half-wavelength dipole in a free-space medium, the radiation resistance becomes approximately 73 ohms.

Secondary Antenna - Parabolic Cylindrical Reflector

A reflector can be used to improve the overall radiation characteristics of the dipole antenna as a primary antenna. A parabolic cylinder is not only the best fit for the dipole antenna, but also mechanically simple to build. As shown in Figure 14, the energy radiated from the dipole is collimated at a line which is parallel to the axis of cylinder through the focal point of reflector. As a result, radiation is very strong and concentrated along the line AB, but cancellation will take place in any other direction, because of path-length differences. Thus the parabolic reflector can radiate a concentrated beam. When the parabolic cylinder antenna is used for reception, waves arriving from the BA direction are brought together at the focus. While waves from any other direction are canceled at that point because of path-length differences.

Since the parabolic cylinder reflector converts the omnidirectional beam coming from the dipole antenna into a concentrated beam in the direction of E-plane in the dipole antenna, which improves the overall performance of antenna radiation and reception characteristics.

Shape of Parabolic Cylinder

The ability of an antenna to concentrate radiation into a narrow beam is

Figure 14. Transmitting and receiving characteristics of microwave beam in the parabola.

limited by the diffraction effect, governed by the relative size of radiating aperture in terms of wavelengths. Therefore, the finite beam width and the spillover effect must be considered in determining the height and length of the parabolic cylinder. The beam width in the plane containing the dipole (horizontal plane) is determined by the radiation of the line source, while the beam width in the perpendicular plane (vertical plane) is determined by the radiation across parabolic profile. Also, if the length of parabolic cylinder is not much longer than the length of dipole, the radiated beam in the horizontal plane can be spilled and diffracted on the edge of parabolic cylinder in the longitudinal direction, which prevents the antenna from radiating a concentrated beam.

The parabolic cylinder was designed based on the analysis of parabolic dish antenna. The angular half-beam width, θ_0 , depends on the wavelength and the antenna aperture diameter (Collin, 1985b):

$$\theta_0 = \frac{70^\circ \cdot \lambda_0}{d} \quad (31)$$

where: d is the diameter of the aperture, and
 λ_0 is the wavelength.

Also, the size of illuminated area on the soil surface is determined by the half-power beam width.

$$A = \pi \cdot \left[h \cdot \tan\left(\frac{\theta_0}{2}\right) \right]^2 \quad (32)$$

where: A is the illuminated area on the soil surface
 h is the distance from antenna to soil surface.

If the antenna was mounted approximately 137 cm above soil surface and the desired illuminated area on soil surface was assumed to be about 79 cm in diameter, the half-beam width from equation (32) became 32° . Substituting the half-beam width into equation (31), the diameter of antenna aperture in the parabolic antenna can be determined. The aperture diameter of 66 cm was used in this study.

While the aperture of parabolic antenna is circular shape, the aperture of parabolic cylinder antenna is rectangular shape. Thus, the height of the

parabolic cylinder was set to 66 cm and the length of cylinder was determined as 52 cm to achieve the equivalent aperture in the parabolic antenna.

When the dipole is used as a feed source for the reflector-type antenna, the rear feed configuration is usually used (Mivec, 1963). Since the rear feed method is mechanically compact and needs a minimum length of transmission line, the effect of aperture blocking and impedance mismatch by the feed itself can be reduced. On the other hand, the ratio of the focal length to the diameter of antenna aperture, (f/D) , is another important parameter in designing parabolic reflector. Small f/D ratio requires a deep reflector, which allows relatively small aperture to be equivalent to the large aperture with the large f/D ratio. Also, the small f/D ratio makes it possible to use a short semi-rigid coaxial cable to feed the dipole, which is suitable for the rear feed configuration. However, this semi-rigid coaxial cable should be at least longer than the quarter-wavelength to provide a space for the impedance matching and the balancing. Thus, the focal length of parabola was set to 11.66 cm as a compromise between the requirements of small f/D ratio and available space. Then the parabola profile, as shown in Figure 15, can be defined by the following equation:

$$y^2 = 4 \cdot (11.66) \cdot x \quad (33)$$

where: all units are centimeter.

Figure 15. Parabolic profile for the parabolic cylinder construction.

For the height of parabolic cylinder of 66 cm, the intercept point on x-axis, or the depth of parabolic cylinder became 23.3 cm.

Four steel sheet-metal were cut into parabola shape according to equation (33) and arranged to make a rib of the parabolic cylinder as shown in Figure 13. For the reflecting surface, very thin aluminum sheet-metal was pushed and glued to the structure of parabolic cylinder rib. Fiberglass resin with mat material was used to reinforce the reflecting surface attached to the parabolic cylinder rib.

Impedance Matching and Balancing between Antenna-Transmission Line

In constructing the antenna, careful attention was given to the antenna balancing and impedance matching. Since the inner and outer conductors of the coaxial cable are not coupled to the antenna in the same way, the coaxial cable is inherently unbalanced. The unbalanced mode currents coupled to the receiver input result in reduction of both gain and signal-to-noise ratio.

As a coupling-matching method, a quarter-wavelength coaxial balun was used in this project. The balun is a device which is used to couple a balanced system to an unbalanced system by canceling or choking the outside current (Collin, 1985a). Two slots were made on the outside conductor of coaxial cable. These slots are little longer than the quarter-wavelength and located on the exact opposite sides of the outside conductor of coaxial cable. The remaining opposite parts of the outer conductor of coax represent two electrically-shortened parallel transmission line. A short length of metal sleeve was placed on the coaxial cable to empirically match the input impedance of the antenna to the radar system. The configuration of the quarter-wavelength balun with the metal sleeve is shown in Figure 16. Using the network analyzer the impedance matching was achieved by moving the metal sleeve up and down on the slotted coaxial cable. Every effort was made to match the impedances of both antennas as close as possible to the system characteristic impedance of 50 ohms at the operating frequency of 1.1 GHz.

Since the radar system actually used two different frequency signals, the impedance matching were necessary at the two discrete frequencies of 0.98 GHz and

Figure 16. Configuration of the quarter-wavelength balun for impedance matching.

1.12 GHz. The impedance mismatch results in the power reflection back to the radar system when the system radiates energy into air. The power reflection is calculated from the reflection coefficient Γ in equation (2.3), where Z_1 is the characteristic impedance of transmission line (50 ohms) and Z_2 the antenna input impedance. The impedance matching is usually considered acceptable if the reflection coefficient Γ is less than 0.2, or the power reflection coefficient Γ^2 is less than 0.04 (Collin, 1985a). The constructed antennas showed an acceptable impedance matching at the frequency of 1.22 GHz and a marginal impedance matching at 0.98 GHz as shown in Table 4

Gain Characteristic of the Antenna

The gain of the antenna was measured by the transmission method (Cutler et al., 1947), which requires two identical antennas. Two identical parabolic cylinder antennas were constructed and placed a distance r apart as shown in Figure 17. The transmission loss at the distance r is defined by the ratio of the received power P_r to the transmitted power P_t as follows:

$$\frac{P_r}{P_t} = \left(\frac{G \cdot \lambda}{4 \cdot \pi \cdot r} \right)^2 \quad (34)$$

where: P_r/P_t is the transmission loss,

G is the antenna gain and

r is the distance between two antennas.

From equation (34), the gain of antenna can be determined by measuring the transmission loss at various distances. A power meter (Hewlett Packard, Model 435B meter & 8481A power sensor) was used to measure the transmitted and the received power. The transmission losses were determined at the distances ranged between 190 cm and 450 cm as shown in Figure 18.

The average gain of the antenna was determined as 10.6 dB. Considering that the gain of ideal dipole was only 2.2 dB from equation (28), the parabolic cylinder reflector greatly improved the system performance.

Bistatic Mode Operation

Bistatic vs. Monostatic Mode

Bistatic mode is a radar operating mode which uses separated transmitting

Table 4.7. Input impedance and reflection coefficient of antennas.

	Transmitting		Receiving	
	Impedance	Γ	Impedance	Γ
@ 0.98 GHz	$20 - j1.2$	0.43	$22 - j4.6$	0.39
@ 1.1 GHz	$49.1 + j34.9$	0.33	$50 + j32$	0.31
@ 1.22 GHz	$68 + j14.4$	0.19	$58 + j5$	0.09

Figure 17. Gain measurement by the transmission method.

and receiving antennas.

Since the dual frequency continuous-wave radar essentially involves the use of two continuous-wave signals, high isolation between the transmitting and receiving functions is required. A circulator (RYT Industries, 300007; 20 dB isolation) which allows a single antenna to be used for the transmitting and receiving functions, was tried first but did not provide sufficient isolation for the system developed in this project. A preliminary experiment showed that at least 40 dB isolation is needed for the system. Therefore, separate antennas were used for transmitting and receiving functions to achieve the high isolation.

When the radar system was designed, the theoretical maximum unambiguous range was 31.25 cm from equation (4.7). While the elapsed time for round-trip of radiated signal in monostatic mode is simply twice as the distance from the antenna to the target, the elapsed time in the bistatic mode is changed according to the bistatic angle and the distance between two antennas. That means the unambiguous range also varies according to the antenna arrangements. As shown in Figure 19, when two antennas are arranged symmetrically and the distance between antennas L and the bistatic angle β , then the elapsed times for round-trip in monostatic and bistatic modes can be expressed as follows:

$$T_m = \frac{2 \cdot F}{c} \quad (35a)$$

$$T_b = (T_t + T_r) = \frac{\frac{2 \cdot F}{c}}{\cos \frac{\beta}{2}} \quad (35b)$$

where: F : the focal length made by two antennas,
 T_m : the elapsed time in monostatic mode,
 T_b : the elapsed time in bistatic mode,
 T_t : the elapsed time in the transmitting path and
 T_r : the elapsed time in the receiving path.

Hence, the unambiguous range in bistatic mode becomes

$$R_b = \frac{R_m}{\cos \frac{\beta}{2}} \quad (36)$$

where: R_m : the unambiguous range in monostatic mode

R_b : the unambiguous range in bistatic mode.

Therefore, the operation of the radar system in bistatic mode extends the unambiguous range by the factor of $1/\cos(\beta/2)$.

Effective Illuminated Area

The effective illuminated area on the target surface is determined by the antenna beam width. In bistatic mode of operation, the analysis of the effective illuminated area becomes much more complicated than in monostatic mode because both the transmitting and the receiving antennas should be focused at the target. Figure 20 illustrates how the overlapping of those illuminated areas on the flat target surface is affected by the focal length of two antennas. The radar system is not able to detect the target surface properly if the operating range of two antennas is excessively out of the focus.

Figure 18. Effect of focusing antennas' beams on the illuminated area.

To determine the effective illuminated area, the effective antenna beam width of two antennas was investigated experimentally. The experimental procedure for measuring the effective antenna beam width was very similar to the method used in the previous chapter for GPR.

Two antennas were located 137 cm above concrete floor. From the imaginary center point made by two antennas, the power of reflected signal on the galvanized sheet metal of 66 cm x 89 cm placed on the concrete floor was measured in the increment of the angular displacement of 4°. The power of the returning signal was reduced as the galvanized sheet metal was moved outward from the center line. After a certain angular displacement, the returning signal is contributed by only the concrete floor, not by the galvanized sheet metal at all. This critical point is determined as the effective antenna beam width. Since the effective illuminated area is affected by the arrangement of two antennas in terms of the bistatic angle and the distance between two antennas, the antenna beam width was measured in both longitudinal and latitudinal directions.

Figure 21 shows the power of reflected signal on the piece of galvanized sheet metal plotted as the angular displacement of antenna beam width. The effective antenna half-beam width of two antennas arrangement in the longitudinal direction was approximately 12°. The effective antenna beam widths in the latitudinal direction were approximately 20° each for the transmitting and the receiving antenna sides. The effective illuminated area A_e can be expressed as

$$A_e = 2 \cdot h^2 \cdot \tan\theta_L \cdot (\tan\theta_{Lr} + \tan\theta_{Lr}) \quad (37)$$

Where, h is the distance between the antenna and target surface, θ_L is the antenna half-beam width in the longitudinal direction, and θ_{Lr} and θ_{Lr} are the antenna beam widths for the transmitting and the receiving antenna sides in the latitudinal direction. For the height of 137 cm, the effective illuminated area became 60 cm x 104 cm.

(a)

(b)

Figure 19. Determination of antenna gain.

EVALUATION OF DEVELOPED RADAR SYSTEM PERFORMANCE

Experimental Setup and Procedure

The prototype dual frequency continuous-wave radar, as shown in Figure 22, was constructed in-house according to the design specifications developed in the previous chapter. The relationship between phase detector output and the distance from the antennas to the target surface were obtained in the laboratory setup and in the agricultural field. Since the phase detector outputs in the developed system seemed to be dependent on target materials, separate experiments were conducted on several target surfaces including the concrete floor, the galvanized sheet metal and various soil surfaces.

A steel structure for mounting two-antennas was implemented on a hand crane, as shown in Figure 23. The bistatic angle and the distance between two antennas were adjustable, and the antennas can be moved up and down over 66 cm range. The lowest point of transmitting antenna can be placed at 76 cm above the target surface. Since the output voltage of phase detector was dependent on the amplitude variation as well as the phase difference of the reflected signals, the arrangement of two antennas was critical to the performance of phase detector. The optimum distance between two antennas and the incidence angles of the transmitting and the receiving antennas were sought by trial and error. They were determined as 114 cm for the distance between two antennas and 19° and 17.5° for the incidence angles of the transmitting and receiving antennas, respectively. To prevent any direct transmission between two antennas, the receiving antenna was located higher than the transmitting antenna by 10 cm.

The experiments were conducted on the concrete floor and on a galvanized sheet metal of 120 cm x 120 cm placed on the concrete floor in the laboratory. The phase detector outputs were measured over the range of 66 cm by moving the two antennas up and down in the increment of 1 cm.

To investigate the dynamic response of developed radar system, two antennas were placed on a moving table of a soil bin and operated horizontally toward a stationary vertical flat surface target. Figure 24 shows the overall view of experimental setup in the laboratory. Galvanized sheet metal of 120 cm x 120 cm

Figure 20. Arrangement of antennas in bistatic mode.

Figure 22. Algorithm for determining the distance to target with the dual-frequency continuous-wave radar system.

Figure 23. The constructed dual-frequency continuous-wave radar system.

was used as the vertical target. The varying height of soil surface was simulated by moving the table back and forth toward the vertical target. The actual distances from the antennas to the target surface were tracked by a displacement transducer, which is a photo sensor mounted on the moving table of soil bin. The transducer produces square-wave signal on every 1 cm along the fixed guide rail in the soil bin. Dynamic measurements were conducted by moving the antenna continuously toward the target surface and back over the same interval where the static measurements were made.

The same experiments were conducted on several soil conditions with different roughness in the agricultural field. The smooth soil surface was plowed and disk-harrowed and the rough soil surface was only plowed. Another experiment was conducted on soil surface covered with grass. The selected experimental plots in the agricultural field are shown in Figure 25. These experiments were conducted for three days in the experimental farm at Clemson University.

Experimental Results

Galvanized Sheet Metal and Concrete Floor

The experimental results on the concrete floor and the galvanized sheet metal are shown in Figure 26.

Although the phase detector output on the concrete floor showed a non-linear relationship between the distance and the phase difference, the radar system can measure the distance between 96 cm and 122 cm since the output increased monotonically in this range.

The galvanized sheet metal surface showed similar relationship between 96 cm and 126 cm except between 99 cm and 104 cm, where the phase detector output did not increase much. Except this range the radar system was able to measure the distance to target surface in 1 cm resolution.

The dynamic measurement results agreed with the static measurement results with no apparent time delay.

It was also observed that, although the double-balanced mixer used as the phase detector is able to produce up to ± 300 mV voltage in ideal condition, the maximum outputs on the concrete floor and the galvanized sheet metal were only

(a)

(b)

Figure 4.15. Experimental sites: (a) smooth soil surface, (b) rough soil surface and (c) grass-covered soil surface.

Figure 29. The phase detector output on grass-covered soil surface.

Figure 28. The phase detector output on concrete floor and on galvanized sheet metal.

± 200 mV and ± 270 mV, respectively. These facts indicated that the phase detector delivered an insufficient output due to the low input level. The average powers of reflected signals on the concrete floor and the galvanized sheet metal at the input side of the phase detector were 2.89 mW and 5.86 mW when the average power of reference signal at the other input side of the phase detector was 2.03 mW. Since the phase detector output varies according to the characteristic of target, separate calibrations on the different target surfaces are required.

Soil Surface

Smooth Bare-Soil Surface

The first experiment on the smooth bare soil surface was conducted on April 28, 1992. The second and third experiments were conducted at a different location in the morning and afternoon of April 30, 1992, respectively. Each experiment was repeated three times without moving the experimental device. Experimental results are shown in Figure 27. Compared with the output voltage on the concrete floor and the galvanized sheet metal, the phase detector output on the smooth soil surface was relatively small. This indicates that the conductivity of bare soil surface was smaller than those of the concrete floor and the galvanized sheet metal. Therefore, the returning signal from the bare soil surface was not so strong to fully drive the phase detector.

The phase detector output on the smooth bare soil surface also varied according to soil condition. The second experiment was conducted after it had rained in the early morning. Compared with the first experiment, it provided the higher output voltage. The reason was that the increased soil moisture content due to rain increased the conductivity of soil, which increased the strength of reflected signal from the soil surface. On the other hand, the third experiment conducted in the afternoon of the same day showed very poor experimental result. Sunny and windy weather condition in the afternoon accelerated the evaporation of moisture on the soil surface and lowered the conductivity of soil surface, which provided insufficient amount of reflection power to the phase detector in the radar system.

Figure 29. The phase detector output on smooth-bare soil surface.

The average powers of returning signals at the input side of the phase detector on the three experiments were 2.54 mW, 2.98 mW and 2.23 mW when the maximum outputs of the phase detector were ± 170 mV, ± 220 mV and ± 170 mV. These facts indicated that the output of the phase detector was affected by the power of returning signal from the target surface.

Rough Bare-Soil Surface

As shown in Figure 28, any relationship between the phase detector output and the distance to target was not observed on the rough soil surface. Due to the severe roughness of target surface, the radiated energy from the transmitting antenna must have been dispersed on the rough soil surface. Therefore, the receiving antenna could not gather the sufficient amount of energy to operate the phase detector properly.

The average power of received signal at the input side of the phase detector was only 0.51 mW.

Soil Surface Covered with Grass

Experimental results on the grass-covered soil surface are shown in Figure 29. Compared with other two experimental results on the bare soil surface, the experiment on the soil surface covered by grass showed good result. Since this experimental plot is relatively flat and highly compacted, the strong signal was reflected on the soil surface. In addition, the overall conductivity of grass-covered soil surface was higher than the one of bare-soil surface, because the grass root contains moisture in subsurface and they prevents the moisture in soil surface from evaporating easily.

The average power of reflected signal at the input side of the phase detector was 3.57 mW and the maximum outputs of the phase detector reached to ± 250 mV.

System Calibration

Before the developed radar system operates on any target surface, the radar system needs to be calibrated for the target surface. The calibration should be conducted on the desired target surface without any vegetation materials. Measuring the phase detector outputs in the 1 cm increment over the full range,

Figure 30. The phase detector output on rough-bare soil surface.

the measuring window is determined at the distances where the phase detector output becomes the maximum and the minimum.

Based on the calibration result a look up table is made. The look up table is a data set of distances which are matched one on one to the phase detector outputs of the radar system. Compared with the look up table loaded in the microcomputer memory, the phase detector output of the radar system is converted into the distance between the antenna and the target surface. This procedure is presented in Figure 30. The computer program for determining the distance to target surface was written in C language .

Evaluation of the System Performance

In the Laboratory

To investigate the effect of vegetation on the reflected signal from the target surface, an experiment was conducted in the laboratory with the grass hay. The experimental units were prepared by evenly distributing and packing hay material into a wooden frame of 122 cm x 305 cm x 12.7 cm, which was large enough to cover the effective illuminated area of the radar system. Total six experimental units were made for three thickness levels with two replications. The density of hay material used in the experiment was 0.024 g/cm³.

The experiment was conducted according to the completely randomized and split-split plot design with three thickness levels of hay materials as whole plots and three moisture content levels as subplots. Three thickness levels of hay material were 12.7 cm, 6.3 cm and 3.2 cm. Three different the moisture content of hay material were presented as MC1, MC2 and MC3. The phase detector outputs were measured at 11 points in the measuring window, which were treated as sub-subplots. The system calibration was conducted on the concrete floor without any hay material and the look up table was made in Table 5. The measuring window was 37 cm interval between 93 cm and 130 cm.

On the randomly selected thickness level of hay material the phase detector outputs were measured at 11 points in the measuring window. After increasing the moisture content level to MC2 by spraying an amount of water evenly, the phase detector outputs were measured at the same points. For the moisture content level MC3, the same amount of water was sprayed on the

hay material again and the measurement of phase detector outputs was repeated. The amounts of water sprayed were presented in Table 6. The same experimental procedure was applied to the remaining 5 experimental units.

Based on the calibration result, the phase detector outputs were converted into the distances. During the conversion process, however, some distance data at the thickness level 12.7 cm with the moisture content levels MC2 and MC3 were out of the calibration range. In this case, the maximum or minimum distances in the measuring window were replaced with those unavailable values.

The measurement error was defined as the difference between the actual distance and the distance measured by the radar system. The statistical analysis showed the three way interaction of measuring points $\times$ moisture content $\times$ thickness to be highly significant, which required an investigation of the simple effects among the subclasses. Further analysis were conducted by comparing the differences between or among levels of the measuring points within each level of the moisture content and the thickness, the moisture content within each level of the measuring points and the thickness, and the thickness within each level of the measuring points and the thickness.

The measurement errors among the measuring points were observed to be highly significant within any levels of the moisture content and the thickness. The means of the measurement errors at each measuring point indicated that the measurement error increased as the measuring points approached to the both ends of the measuring window. In other words, the measurement errors were very small in the neighborhood of phase difference of 90° . Since the relationship between phase detector output and the distance to target was linear with the steep slope in the middle range of the measuring window, the measurement error became small even when the phase detector output was strongly affected by the amplitude variation of the reflected signal.

The effect of moisture content on the measurement error was shown in Figure 31. Although the significant difference among the moisture content levels was observed only at 6.3 cm thickness of hay material, the overall measurement error increased as the moisture content increased. Since the conductivity of hay material increased as the moisture content increased, the attenuation due to vegetation material increased as well. Not only the reflected signal on true

Table 4.9. The amount of water sprayed on the hay material.

Moisture content level	Amount of water(ml)		
	3.2 cm	6.3 cm	12.7 cm
MC1	0	0	0
MC2	500	1000	2000
MC3	1000	2000	4000

target surface became weak due to the attenuation, but also the reflection on the vegetation surface possibly dominated the reflection on the true target surface, which resulted in causing the measurement error.

The effect of hay material thickness on the measurement error was shown in Figure 32. It was shown that the measurement errors among the thickness were highly significant different at the moisture level MC3 and the most of measuring points. As the moisture content decreased, the significant difference among the thickness disappeared gradually. This fact indicated that the measurement errors were not affected by the thickness levels of hay materials at the relatively low moisture content.

In the Agricultural Field

For the evaluation of the developed radar system performance, a field experiment was conducted on the same barley field in Simpson Experimental Station, Pendleton, SC, where the performance of ground-penetrating radar has been evaluated. The barley was almost full-grown and the average height of barley was 91.5 cm.

To investigate the effect of crop on the soil surface return of microwave, the experiment was conducted according to the split-plot design. Four different heights of barley were treated as whole plots and the eight points measurement in the measuring window as subplots. The four different heights of barley were the full height of standing crop, after cutting off the grain-head, after cutting the remaining barley stem in half and after removing barley completely. The average height of standing barley in the whole plot were 91.5 cm, 76.2 cm, 38 cm and 1 cm, respectively. The entire experiment was replicated on the two experimental spots R1 and R2 in the barley field. The size of experimental spot was 122 cm x 305 cm, which was large enough to avoid the border effect.

The system calibration was conducted on the bare soil surface where the existing barley was completely removed within the range of effective illuminated area. Then, the phase detector outputs were measured over the barley.

As shown in Figure 33, the measuring window was determined at 99 cm and 133 cm from the calibration result. The experimental results were shown in Figure 34.

Figure 32. The measurement errors among the moisture content levels on the fixed thickness levels of (a) 3.2 cm, (b) 6.3 cm, and (c) 12.7 cm.

Figure 35. The measurements errors among different heights of barley on the experimental spots (a) R1 and (b) R2.

Figure 34. Calibration result of the radar system in the barley field.

The measurement errors according to the barley height variations were not significantly different at 95% confidence level, which were less than 2 cm in the most cases. This analysis indicated that the attenuation and the reflection due to the vegetation did not play an important role in the microwave reflection on true soil surface. It also showed that there were significant difference among the measurement points at 95% confidence level. The measurement error increased as the measurement points approached to the both ends of the measuring window. The maximum measurement errors were 8.91 cm and 8.44 cm for the experimental spots R1 and R2, which were commonly found at the either end of the measuring window.

The field experimental results showed that the performance of developed radar system was mainly affected by the soil surface rather than the vegetation material. In other words, the radar system was able to penetrate through the vegetation material and detect soil surface. Electronic components for a system similar to the US dual frequency continuous-wave were purchase and shipped to the Agricultural Engineering Department. Antenna were made in Israel. Testing of the unit gave very similar results as to the one in the US. However the sensitive of the system was not as good as the one in the US. The reasons for this could be readily determined. The system remains available for further testing as resources are found.

CONCLUSIONS

During the course of the project, three different radar technologies were investigated. All have limitations that made them unsuitable for the input into an automatic controller for agricultural machinery. The ground penetrating radar has the ability to penetrate the plant foliage and reflect a reasonable signal to the antenna. However, the system operates in a pulse mode and the time between pulses is too long for accurate control.

The manufacturer was contacted about modifying the unit for our use. The manufacturer expressed no interest in any cooperation and we would have to use the equipment as delivered. An investigation into a local modification of the equipment was made. Due to the high level of complexity with the signal processing, it was determined that without the support of the manufacturer no modification would be possible. This is a technology that is possible provided a complete radar testing laboratory would be available with a long term commitment and a large operating budget. The investigators were unable to convince Geo-Physical Survey, Inc that a viable market exists for such a device. Their traditional market for GPR systems has been low volume-high price. The idea of a low cost-high volume market could not be accepted. There have been new companies moving into the GPR. If they could be interested in such a project, we could see advances in this technology.

The radar system by Elta has severe limitation. Operating at 2.2 GHz, the effect of dense plant material is large and the system is not able to distinguish between the top of the plant canopy and the soil surface. Here again the manufacturer is not willing to cooperate with the investigators. This technology is one that simply does not work. If other frequencies could be used coupled with the high level of expertise in the company, the outcome could be different.

The dual-frequency continuous-wave gave the most promising results. However the effects of the plant foliage was more than expected. The conditions at the soil surface such as moisture and roughness had a strong influence on the output. The system as tested would require a calibration for each field condition. This system was not acceptable for further testing without serious improvements to the components.

Although no completely satisfactory system was developed, this project has laid the preliminary ground for future studies. As the world of electronics is rapidly changing, new and more precise components could lead to new dimension in automatic control of agricultural machinery.

DESCRIPTION OF COOPERATION

There has been excellent cooperation between the investigators. The e-mail system has proved itself many times in both the preparation of the proposal and execution of the research. At points in the research, daily messages were being exchanged with the investigators.

While Dr. Schmulvich was on sabbatical leave at Texas A&M, several meetings were held between himself and the US investigators. Two trip were made to the US by Dr. Peiper during the project. Dr. Dodd made one trip to Israel to work with both the system at the Agricultural Engineering Department at Volcani and the system at the Agricultural Engineering Department at the Technion.

EVALUATION OF THE RESEARCH

Objective 1 was completed for all three systems tested. Preliminary tests were completed which indicated there was potential use of any of the system. To comply with objective 2, extensive testing was conducted for each system under different field conditions. This is where definite problems started to arise. Each system has limitations that would prevent it from use on agricultural machinery. The work has opened a new area of research in sensor development. Although we did not find a completely acceptable system, we have explored and documented the response of three different types of radar signals. Objective 3 was not reached because an acceptable sensing system could not be developed for incorporation into the control system.

The benefit to agriculture remains extremely high. Improved efficiencies for harvesting machines would add dramatically to the profit margin for the growers. Harvesting losses are extremely expensive because resources to produce the commodity has been spent and considerable labor charges have been incurred only to leave the mature material in the field.

Project Publication

Shin B., R. B. Dodd and Y. J. Han. 1991 Ground Surface Detection Using Dual Frequency Continuous-Wave Radar. ASAE paper 91-3450, St Joseph, MI, USA

Shin, B., R. B. Dodd and Y. J. Han. 1993 Dual Frequency Continuous-Wave Radar for Ground Surface Detection, Trans. Of ASAE, In press