

The "Current Driven Model" -- Experimental Verification and the Contribution of I_{dd} Delta to Digital Device Radiation

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Abstract: Several researchers have proposed that a primary source of emissions from digital devices is due to the partial inductance of the return trace on printed circuit boards. In this "current driven model," RF currents derived from the nanosecond rise time of periodic signals such as clocks create a voltage across the return due to this inductance. This paper reports on an experimental verification of this model, but points out apparent limitations -- at frequencies above a certain point, internal characteristics of integrated circuits such as I_{dd} Delta appear to be the dominate cause of emissions, at least in the circuits examined.

INTRODUCTION

Within the last decade, a number of models have been proposed to explain why digital devices radiate. In this paper, we review these models and perform some tests to check whether a simple circuit produces the emissions theory predicts.

One such theory is known as the "current driven mechanism" and is illustrated in Figure 1. A clock or clock/driver combination serves as a source driving a distant load. The signal produced is a trapezoidal wave and the source has an internal resistance, R_o , and inductance, L_o . The load is a logic gate, which, for MOS based technologies, can be modeled as a capacitance. A series resistance is sometimes inserted at the source end to suppress ringing.

The trace connecting the source to the load, the theory predicts, can be characterized as an inductor. Similarly, the return plane can also be characterized as an inductor at the frequencies of interest.

If we know the current passing through the return plane, then using the inductance the models predict we can calculate a voltage drop across the return plane. This voltage drop will drive any wires attached to the return system as if they were antennas. Basically, the return plane serves as a low impedance voltage source driving attached wires. Any wires directly or incremental connected to the return plane will radiate. In a worst case scenario, the wires attached to the return plane can be stretched out to form a dipole resonant at one of the harmonics at the clock oscillator.

In order to evaluate the theory's predictions, we built the circuit in Figure 2. A 6¼" by 4½" (15.9 cm by 11.4 cm) printed circuit board was constructed with the top side of the board reserved for a V+ plane and the bottom side a V-. A clock oscillator, an Epson SG51P, was placed to one side of the board spaced 3.75" (9.5 cm) away from a 74HC02 device serving as a load. The power supply consisted of a

9 volt battery feeding a 7805 regulator whose output was loaded with a 10 microfarad tantalum capacitor. Placed beneath the clock oscillator and the 74HC02 device were .02 uF wafer type bypass capacitors from Circuit Components Inc., Part No. 293A14. The board was made out of a phenolic material and was .07" (1.8 mm) thick.

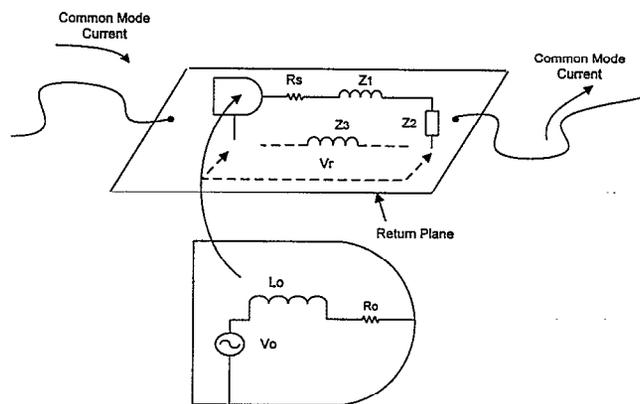


Figure 1

In order to connect the source to the load a wire was used. It was placed adjacent to the underside of the board. The conductor was solid copper .03" (.76 mm) in diameter. Its insulation was .015" (.38 mm) thick. A 50 ohm carbon composition resistor was connected immediately at the output of the clock driver. A 50 picofarad capacitor was placed at the input of the 74HC02 device to simulate heavy loading. The clock had specs typical of a HC device.

In order to simulate the effect of radiation off attached I/O, two telescoping antennas elements were attached to either side of the PCB and were electrically connected only to the V- plane. A Tektronix CT1 current probe was used to measure the current through the wire. At a frequency of 125 MHz the amplitude of the current measured was 2.8 milliamps RMS. The current probe was removed during radiation testing.

Emissions tests were performed at an open field test site. The site has been previously checked against open area test site standards and had been accredited by NIST. Measurements were taken atop a .8 meter

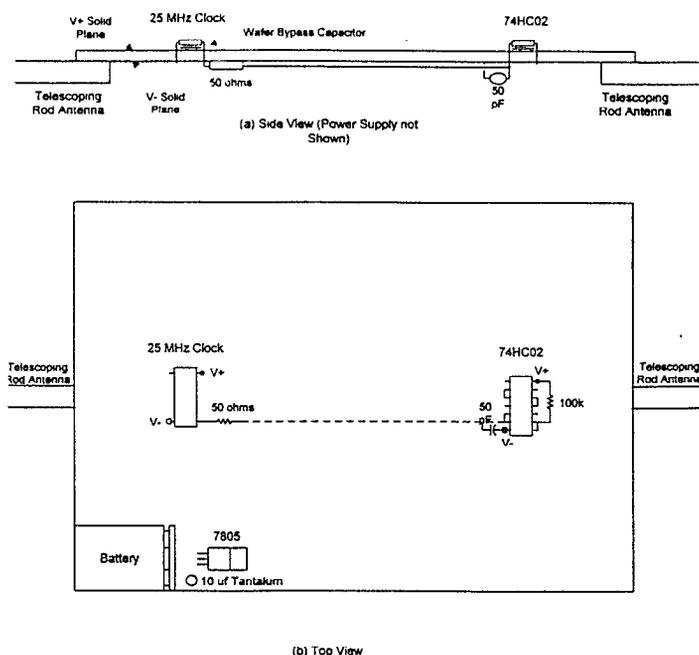


Figure 2

wooden turntable which was rotated to detect maximum emissions. As expected, when the attached telescoping antennas were tuned for resonance, maximum emissions at the resonant frequency were observed when the telescoping arms lay parallel to a horizontally polarized antenna. Measurements were performed at a distance of 10 meters and the antenna was raised and lowered to detect maximum emissions over a 1 to 4 meter range.

“CURRENT DRIVEN” MODEL PREDICTIONS

We began our examination by focusing on one frequency, the fifth harmonic of the clock at 125 MHz. The telescoping elements were tuned to resonance at that frequency and left there for the duration of the test. The circuit shown in Figure 2 produced 39.4 dBuV/m of radiation at 10 meters. The inductance of the return plane, according to Kaden as reported by Leferink [1], is predicted to be:

$$L_{return\ plane} = \frac{\mu l}{2\pi} \left(\frac{d}{w}\right) \approx 2 \left(\frac{d}{w}\right) \text{ nH/cm} \quad (1)$$

Where:

$L_{return\ plane}$ = return plane inductance

w = width of the plane in meters

d = distance between the signal trace and the return plane (meters)

l = length of the trace in meters, $l \gg d$

A slightly different estimate has been made by Hockanson, et al [2]. It is:

$$L_{return\ plane} \approx 5 \left(\frac{d}{w}\right) \text{ nH/cm} \quad (2)$$

We will use the latter formulation to arrive at a prediction for the radiation we should observe. Inserting the values for the circuit in Figure 2 ($d=.76\text{ mm}$, $w=114\text{ mm}$, $l=9.5\text{ cm}$) yields an inductance value for the return plane of .033 nH/cm or .32 nH in total. At 125 MHz an impedance of .25j ohms would result due to this inductance. The voltage drop across the return can be readily computed from the measured current at 125 MHz (2.8 milliamps). The voltage across the return, the model predicts, is .07 Volts.

This voltage drives the attached telescoping antenna, the arms of which were adjusted to half wave resonance. We can calculate the predicted free space emissions using the following formula:

$$E(V/m) \approx \frac{7}{10} \frac{V_r}{\sqrt{3}} = .082 V_r \quad (3)$$

Where:

$E(V/m)$ = Free space field strength at 10 meters

V_r = Voltage across the return trace

Our model predicts a free space radiation level of 35.2 dBuV/m at 10 meters.

Testing over a ground plane affects the impedance of the circuit somewhat and provides for ground reflection. As an approximation, we can assume that the net of these effects is to increase emissions by 5 dB at 125 MHz. Using this adjustment, our model predicts emissions of 40.2 dBuV/m, quite close to the measured value. We can conclude that the model reasonably predicts emissions from the circuit of Figure 2.

IMPERFECT POWER PLANES

The circuit of Figure 2 used solid power planes. Practical power planes, however, are not solid but are interrupted by holes and gaps. Models proposed by researchers predict that emissions will rise dramatically if the return plane is interrupted with a slit as shown in Figure 3a. The slit cuts completely through the PCB, interrupting both the V+ and V- planes. It is .065" (1.65 mm) wide and extends from one edge of the board to a point 1" (2.54 cm) past the trace. The measured emissions at 125 MHz did rise dramatically, to 59.8 dBuV/m.

Hill, et al., [3] models the increased inductance by analyzing the gap as a shorted transmission line. The Appendix uses this theory to derive the impedance of a slotted return plane. It is predicted to be:

$$L(\text{gap thru } V+, V-) \approx \frac{10}{2 + \ln\left(\frac{w}{s}\right)} \text{ nH/cm} \quad (4)$$

Where:

w = the width of the plane to the left and right of the slot in meters

s = the width of the slot itself in meters

$w \gg s$ and $l_{gap} \ll \lambda$

Applying this formula to our test circuit ($s=1.65\text{ mm}$, $w=2.7"/6.86\text{ cm}$, $l_{gap}=2.54\text{ cm}$) yields a predicted value of return plane inductance of 4.4 nH resulting in predicted emissions of 63.0 dBuV/m at 10 meters. This value is in reasonable agreement with the measured value.

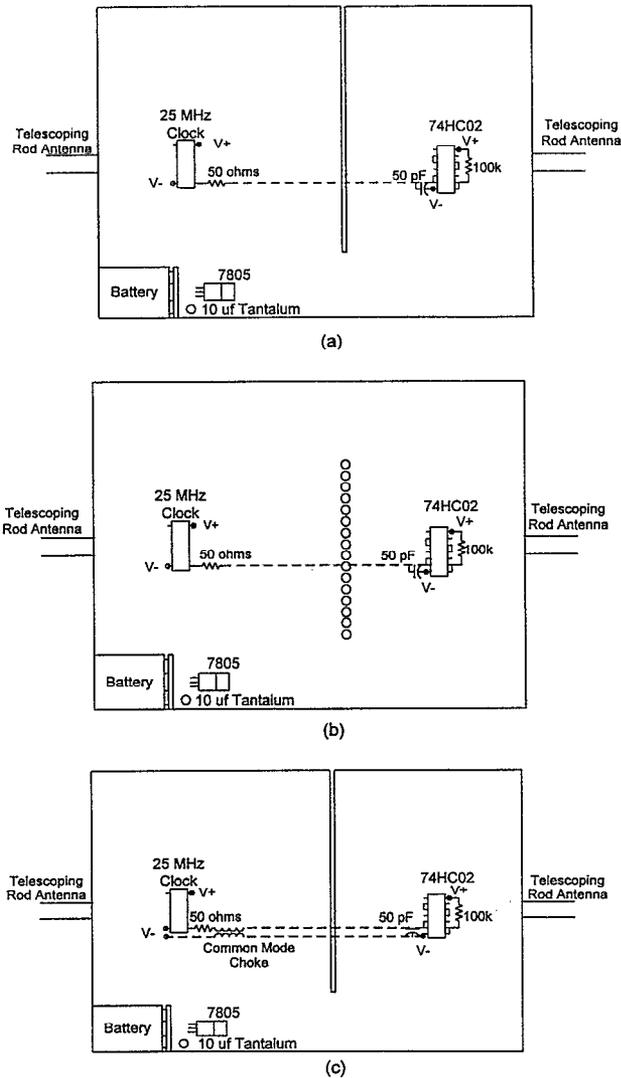


Figure 3

Researchers also agree that if the return plane is interrupted by holes rather than a slit, the increased inductance caused by the presence of the holes will increase emissions only slightly. To test this theory, we modified the circuit of Figure 2 by drilling holes through the V+ and V- planes, as shown in Figure 3b. These holes were placed .16" (4.1 mm) center to center and were .125" (3.2 mm) in diameter. No change in emissions was noted at 125 MHz due to the presence of the holes.

Next, we evaluated a different method for reducing emissions from an imperfect return plane. This method uses a common mode choke located near the clock [4]. In theory, the presence of the common mode choke should force current to return through the return wire, the one which passes through the common mode choke, instead of through the return plane. Even if the return plane was inductive because of the presence of an opening such as a slit, little voltage would be dropped across the return plane simply because the RF

current does not pass through it.

We used the circuit of Figure 3c. A twisted pair consisting of 24 AWG magnet wire was passed through two Fair-Rite 2643000801 No. 43 type ferrite beads 1½ times and was then connected the clock and the load. The return wire was connected to the ground plane immediately adjacent to the clock and the load. Emissions fell dramatically at 125 MHz, to 38.7 dBuV/m at 10 meters.

Emissions were then measured using a circuit that employed both a common mode choke, as shown in Figure 3c, and the solid ground plane of Figure 2. See Figure 5a. Emissions fell once again, this time to 32.7 dBuV/m.

As a final test the connection between the clock and the load was removed so that the clock oscillator could run by itself without any wires attached. At 125 MHz the clock oscillator, operating alone and fed power through solid V+ and V- planes, produced 29.7 dBuV/m of emissions, only 3 dB less than the emissions produced by the use of a combination of a common mode choke and a solid return plane.

We can conclude that the models proposed by researchers do a reasonable job of predicting the emissions at the fifth harmonic, 125 MHz. Things were different, however, when it came to the ninth harmonic at 225 MHz. Table 2 tabulates the emission results obtained at that frequency. For this set of tests, the telescoping antennas used to simulate connected I/O cables were left at their 125 MHz resonance length of 19". What remarkable about the data in Table 2 is that it is seemingly unaffected by all the changes that were made. A return plane with a gap in it exhibited much the same radiation as one which used solid return planes. It appears that a different mechanism is at work at these higher frequencies.

Idd DELTA

The integrated circuits we used were of the MOS family. Figure 4 shows the basic structure of a MOS device. A P channel and N channel device serve as switches alternately connecting the output to V+ and V-, depending on the input state. When the input is high, the N channel FET is turned on, connecting the output to V-. The P channel device is in its off state and presents a very high impedance between V+ and the output. Therefore, little current flows between V+ and V-. The same situation is true in reverse when input is low and the output is high. In the transition region, however, current does flow from V+ to V-. This current is a function of input voltage. It peaks in the middle of the input voltage range and is known as I_{dd} Delta.

The affect of I_{dd} Delta is to produce a very brief current pulse every time the gate changes state. If our simplified gate of Figure 4 is being driven by a typical HC signal, (risetime approximately 5 nanoseconds) the input voltage passes briefly through the range where I_{dd} Delta is significant as the signal transits from logical zero (0 volts) to logical one (5 volts). The net result is a current pulse of approximately 1 milliamp peak and about 1 nanosecond in width.

Unfortunately, the amount of radiation we can expect due to the affect of I_{dd} Delta can be difficult to predict. For one thing, manufacturers rarely cite I_{dd} Delta in their data sheets. For another, I_{dd} Delta is highly variable. It is a function of the supply voltage to the gate, varying as a function of V_{cc} to the 2.2 power.

Table 1: Emissions at 125 MHz

Test Condition	Figure	Emissions dBuV/m at 10m
Solid Return Plane	Figure 2	39.4
Slotted Return Plane	Figure 3a	59.8
Holed Return Plane	Figure 3b	40.2
Slotted Return Plane with CM Choke	Figure 3c	38.7
Solid Return Plane with CM Choke	Figure 5	32.7
Clock Running Alone — No Wires Attached	N/A	29.7

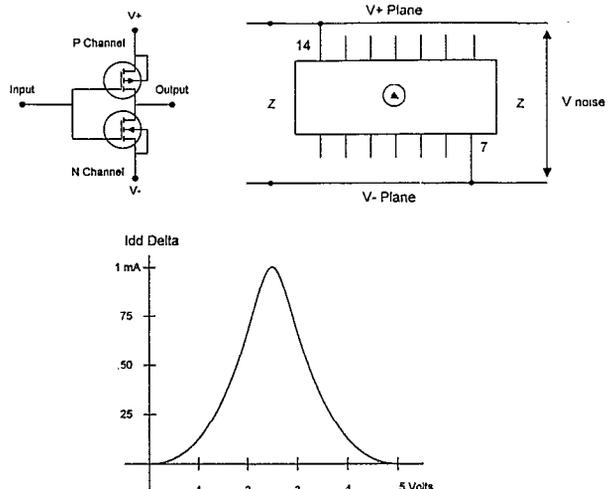


Figure 4

Table 2: Emissions at 225 MHz

Test Condition	Figure	Emissions dBuV/m at 10m
Solid Return Plane	Figure 2	50.2
Slotted Return Plane	Figure 3a	51.2
Holed Return Plane	Figure 3b	50.1
Slotted Return Plane with CM Choke	Figure 3c	49.6
Solid Return Plane with CM Choke	Figure 5	49.6
Clock Running Alone — No Wires Attached	N/A	50.1

I_{dd} current mostly passes through the bypass capacitor immediately adjacent to the integrated circuit. However, the impedance of that capacitor is finite, and some of the current is fed back through the supply planes, distributing it around the circuit. This creates a noise voltage of magnitude $V=IZ$ where Z is the impedance of the supply. This impedance is low, but still significant, and is largely inductive in nature at the higher frequencies.

Some designers attack problems created by the interworkings of integrated circuits by filtering the supply at the IC's supply pins. We

evaluated two strategies for doing so. Our test circuits already had wafer type capacitors placed immediately below them and so we first took the additional step of isolating the V+ pin (pin 14 on both devices) from the V+ plane. A wire 1" (2.54 cm) was used to connect pin 14 on both devices to the V+ plane. Although identical on a schematic, this configuration provided some filtering because of the wire's inductance. Test results show a reduction of 9 dB at 225 MHz. The next step was to add a second bypass capacitor (a 1000 picofarad surface mount multilayer type) and to replace the wire with a surface mount device designed to increase series impedance over a wide frequency range. A TDK device was chosen which, according to the manufacturer's data sheet exhibited greater than 300 ohms of impedance up to 1 GHz. An additional reduction of more than 8 dB was noted.

The data shows that pi type filters are more effective than simple capacitive bypassing in reducing emissions. Some designers, however, prefer not to isolate the supply pin on an integrated circuit (pin 14 in our case) from the V+ plane. To do so will introduce an impedance between the supply pin on the IC and the supply plane which will cause a small voltage drop when the device switches. This voltage drop can reduce noise margins. Nonetheless, we believe that it is prudent to build in pi type filter topography into designs so that it can be used if needed. If testing shows it is not needed, then the series element can be removed.

Table 3 demonstrates the results of our efforts. If we start with our circuit of Figure 3a as a control, we see a dramatic improvement, over 20dB at some frequencies, as modifications were made. The improvement was a function of the use of solid supply planes, a common mode choke between source and load and a pi type filter filtering system at both devices. Note that this improvement was achieved without using any filtering near our "I/O" (telescoping elements) or shielding.

REFERENCES

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- [3] R. Hill, T. Van Doren, T. Hubing, J. Drewniak, and F. Gisin, "Common Mode Currents Induced On Wires Attached To Multilayered Printed Wire Boards With Segmented Ground Planes," *1994 IEEE International Symposium on Electromagnetic Compatibility*, Page 116.
- [4] F. J. Tilley, "Reducing Radiated Emissions on High Speed Signal Lines Using Common Mode Choke Coils," *IEEE Symposium on Electromagnetic Compatibility*, 1995.
- [5] C. R. Paul, *Introduction to Electromagnetic Compatibility*, John Wiley and Sons, 1992, Page 271.

APPENDIX

The derivation of the impedance of a slot in a plane is usually based on the assumption that the slot "looks" like a shorted transmission line to the return current. [3, 5] This is illustrated in Figure 6. The hatched area in the figure represents the area where most of the return current flows. This can be modeled as a coplanar transmission line consisting of two conductors of widths w_1 and w_2 separated by an air gap of width s . The coplanar transmission line is shorted as its end and, in our test circuit, is 2.54 cm in length.

An electrically short, shorted transmission line acts as an inductor. Its inductance is:

$$L(\text{gap}) = Z_0 \frac{\tan B l_{\text{gap}}}{\omega} \quad (5)$$

Where:

$L(\text{gap})$ = the inductance of the gap in Henries

$B = 2\pi/\lambda$

λ = wavelength in the media

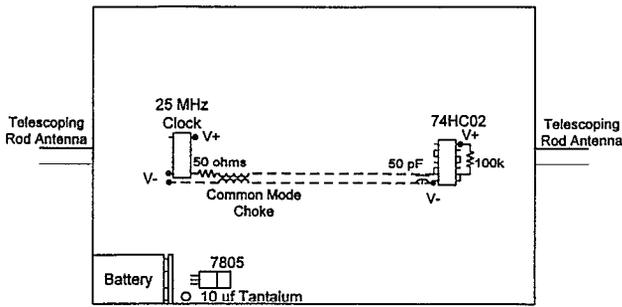
l_{gap} = length of the gap in meters

ω = frequency in radians per second

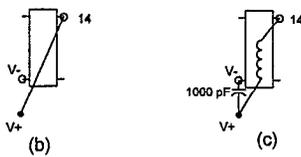
The characteristic impedance of the transmission line is a function of the width of the conductors making up the line. In our case, these are unequal and that is not one of the cases modeled in the references. Therefore, to make things easier we will make the worst case assumption that the two planar conductors making up the strips of the transmission line are equal and of width w_1 . The impedance of such a transmission line is: [3,5]

$$Z_0 = \frac{377}{\sqrt{\epsilon_r}} \frac{K(k)}{K'(k)} \quad k = \frac{s}{s+2w} \quad k' = \sqrt{1-k^2} \quad (6)$$

$$\frac{K(k)}{K'(k)} = \frac{\pi}{\ln\left(2 \frac{1+\sqrt{k'}}{1-\sqrt{k'}}\right)}$$



(a)



Note: Wafer Type Capacitor Under all ICs (.02 uF)

Figure 5

Table 3. Reduction in Emissions (dBuV/m at 10m)

Frequency (MHz)	Circuit of Figure 3a	Circuit of Fig. 5 (with Pi Filter)
75	41.3	27.3
125	59.8	31.2
175	53.4	34.3
225	51.2	33.6
275	33.8	27.8
325	48.4	22.7
375	48.4	<20
425	39.4	<20
475	37.3	<20
525	31.7	<20

CONCLUSIONS

The "current driven model" does indeed appear to predict emissions from digital devices at the first few harmonics of the clock. However, at higher frequencies, the role of I_{dd} Delta appears to become more important and some additional measures, such as a filter at the integrated circuit's supply, appears to be needed to reduce emissions.

Where:

- ϵ_r = relative permittivity
- s = width of the gap in meters
- w = width of the strips in meters

Since in the usual case $w \gg s$, k is much less than 1. That allows us to make some simplifying assumptions:

$$k' \approx 1 - \frac{k^2}{2} \quad \sqrt{k'} \approx 1 - \frac{k^2}{4} \quad 1 - \sqrt{k'} \approx \frac{k^2}{4} \quad (7)$$

$$1 + \sqrt{k'} \approx 2$$

Therefore:

$$Z_0 \approx \frac{377}{\sqrt{\epsilon_r}} \left(\frac{\pi}{\ln \frac{16}{k^2}} \right) \quad (8)$$

Substituting this into the equation for L_{gap} yields:

$$L(gap) \approx \frac{377}{\sqrt{\epsilon_r}} \left(\frac{\pi}{\ln \frac{16}{k^2}} \right) \frac{\tan \beta l_{gap}}{\omega} \quad (9)$$

$$\frac{B}{\omega} = \frac{\sqrt{\epsilon_r}}{c} \quad \beta l_{gap} \ll 1 \quad \frac{\tan \beta l_{gap}}{\omega} \approx \frac{\beta l_{gap}}{\omega}$$

$$L(gap) \approx \frac{377}{c} \frac{l_{gap}}{\ln \frac{16}{k^2}} \approx (39.5 \times 10^{-9}) \frac{l_{gap}}{\ln \frac{16}{k^2}}$$

We can simplify this again by restating the inductance in terms of nanohenries per centimeter of gap length:

$$L(gap) \approx \frac{39.5}{\ln \frac{16}{k^2}} \approx \frac{39.5}{2.77 - \ln k^2} \text{ nH/cm} \quad (10)$$

$$k \approx \frac{s}{2w} \quad \text{since } w \gg s$$

$$L(gap) \approx \frac{39.5}{4.15 + 2 \ln \left(\frac{w}{s} \right)} \approx \frac{20}{2 + \ln \left(\frac{w}{s} \right)} \text{ nH/cm}$$

Finally, we note that in our example the gap cuts through both the V+ and the V- planes. If these planes are considered to be well coupled, then the gaps formed in the two planes will act like two inductors in parallel, having the total return inductance. In this case, the inductance per unit length would be one half that shown in the equation above:

$$L(\text{gap thru } V+, V-) \approx \frac{10}{2 + \ln \left(\frac{w}{s} \right)} \text{ nH/cm} \quad (11)$$

Where:

- $w \gg s$
- wavelength $\gg l_{gap}$

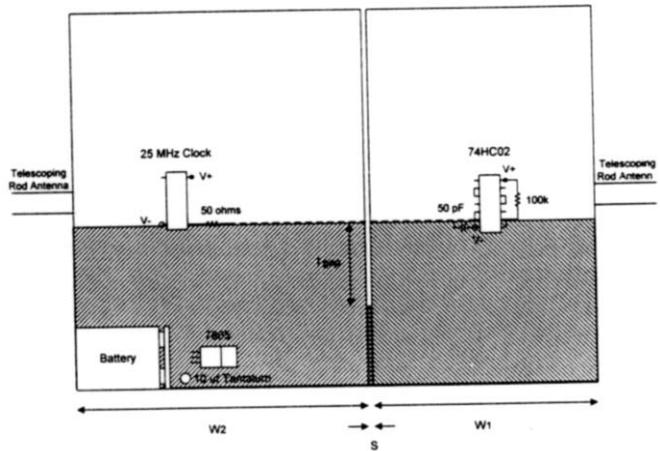


Figure 6