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Master Thesis

für

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Thema: Design of an UWB monostatic microwave radar

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Kennwort: UWB monostatic radar.

Thema: Design of an UWB monostatic microwave radar.

Aufgabenstellung: Based on the realized ultra-wideband (UWB) bistatic radar, a concept of an UWB monostatic radar should be developed. For the system realization, an UWB coupler is needed for the detection of the transmitted and received signals. The device should be designed for ultra-short signals with a pulse width of about 75 ps.
The device should be designed with ADS, fabricated and measured. A demonstrator of an UWB monostatic radar with the new device should then be built up including earlier developed components used for the bistatic radar system.
Further information will be given by the supervisor.

Betreuer: Dipl-Ing. Amnoiy Ruengwaree

Kassel, 15/1/2003

Master Thesis

Ausgeführt von:

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Thema: Design of an UWB monostatic microwave radar

Im Fachgebiet Hochfrequenztechnik
Universität Kassel

Fertiggestellt am
11.07.2003

Hiermit erkläre ich, dass die vorliegende Diplomarbeit von mir selbständig verfasst worden ist, und dass zu deren Anfertigung keine anderen Quellen oder Hilfsmittel hinzugezogen wurden, als die von mir angegebenen.

Kassel, den 11.Juli 2003

(Mohamed El-Hadidy)

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To my parents who I loved too much, and to my son 'Hamza' who I missed too much.

Abstract

In this work a concept of an ultra-wideband (UWB) monostatic radar should be developed based on the realized UWB bistatic radar.

For the system realization, an UWB coupler is needed for the detection of the transmitted and received signals. The device should be designed for ultra-short signals with a pulse width of about 75 ps.

Three port UWB directional coupler and signal divider using the hybrid technology is presented. This work is very necessary for the development of a pulsed radar system for near range measurement.

Three configurations of the device are already designed with ADS, fabricated and measured. A demonstration of an UWB monostatic radar with the new device was built up including earlier developed components used for the bistatic radar system.

The time domain measurement using sampling oscilloscope and the S-parameters measurements using network analyser verified the simulation results for all of the three configurations of this device.

The verification confirmed the coupler ability to cover this UWB range with very good performance.

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Chapter 3 : UWB 3-port directional coupler and signal divider design basis

Chapter 4 : Signal divider configurations

Chapter 1 :

Introduction

Chapter 1: Introduction

1.1 Introduction

This work is about radar systems using wide relative (proportional) bandwidth signals called ultra-wideband (UWB) waveforms. The potential advantages of using UWB waveforms for radar include better spatial resolution, detectable materials penetration, easier target information recovery tram reflected signals, and lower probability of intercept signals than with narrowband signals.

Designing UWB radar systems requires considering what happens when a signal is no longer a single, long duration sinusoidal wave. This presents principles needed for understanding UWB radar concepts and their potential capabilities and provides a basis for further investigation, design, analysis, and fabrication.

Radar systems use radiated and reflected electromagnetic waves to detect, locate, and identify targets. Radar systems is a broad term including everything from small police and ground-probing systems to the large radars for ballistic missile defense and airspace surveillance and tracking.

Radar targets may include ground discontinuities, buried objects, stationary objects for navigation, and moving objects including vehicles from automobiles to reenter vehicle systems. The radar system designer's problem is to balance the user's needs and desires with available technology, achievable performance, and affordable cost. The designer's objective is to satisfy the radar user's needs effectively and cheaply. The radar system user is the final judge of acceptable radar performance and cost.

1.1.1 UWB radar terminology and concepts

UWB terminology and definitions are not standardized as of this writing, and this may cause so me confusion in literature searching. Terms such as narrowband and wideband can have several meanings depending on the subject, i.e., communications, radar, etc. So, we will discuss shortly what UWB radar is generally accepted to be and some of the alternative, but related terminology. Assumptions concerning meanings can be misleading. When in doubt, see how the writer uses the term and determine what is being described in basic functional concepts. The best advice is to be aware that UWB may also be called impulse, time domain, nonsinusoidal, baseband, video pulse, ultrahigh resolution, carrierless, super wideband, and other terms.

1.1.2 UWB defined

The defense advanced research project agency (DARPA) 1990 *Assessment of Ultra- Wideband Radar* advised that "definitions need liberal interpretation and that mathematical definitions are difficult to achieve and seldom useful in a practical sense." The following definitions were given.

Energy Bandwidth, B_E :

The energy bandwidth is the frequency range within which some specified fraction, say 90 or 99%, of the total signal energy lies. This must be defined for a single pulse, if all pulses are the same, or for a group of pulses that are processed together to yield a single decision. The upper limit of this range is denoted here by f_H and the lower limit by f_L.

Time-Bandwidth Product, TB:

The time-bandwidth product of a signal is defined as the product of the energy bandwidth and the effective duration of a single pulse or pulse group. It is the measure of the increase in peak signal-to-noise ratio that can be achieved in the radar receiver by appropriate signal processing [1].

Bandwidth is defined as fractional and relative:

$$\text{Fractional Bandwidth} = \frac{2*(f_H - f_L)}{(f_H + f_L)} \tag{1.1}$$

$$\text{Relative Bandwidth} = \frac{f_H - f_L}{f_H + f_L} \tag{1.2}$$

The DARPA panel accepted the following definition: "Ultra-wideband radar is any radar whose fractional bandwidth is greater than 0.25, regardless of the center frequency or the signal time-bandwidth product."

The term ultra-wideband refers to:

Electromagnetic signal waveforms that have instantaneous fractional bandwidths greater than 0.25 with respect to a center frequency. There is no accepted standard usage for UWB terms; writers also use percent bandwidth and proportional bandwidth instead of fractional bandwidth. There are two other radar classes identified by signal fractional bandwidth: narrowband, where the fractional bandwidth is less than 1 %, and wideband, with a fractional bandwidth from 1 to 25 %. These terms were specifically proposed for describing radar systems in 1989. Some confusion results because narrowband and wideband have very different meanings when describing communications channel bandwidths. Most narrowband systems carry information, also called the baseband signal, as a modulation of a much higher carrier frequency signal.

The important distinction is that the UWB waveform combines the carrier and baseband signal. Baseband or impulse radar (or radio) are other names for UWB radar and radio signals. The UWB signals generally occur as either short duration impulse signals and as nonsinusoidal (e.g., square, triangular, chirped) waveforms.

The rule of thumb is that sinusoidal wave signal bandwidth (BW) for pulse signals are inversely proportional to pulse duration (τ), or $BW = 1/\tau$. When the duration of a short sine wave pulse signal approaches several periods, then the relative bandwidth starts becoming a larger fractional value. There are also long duration nonsinusoidal waveforms having significant power at multiples of its fundamental frequency.

1.1.3 UWB terminology and usage

The term UWB is a new term associated with radio and radar technologies called impulse, nonsinusoidal, baseband, video pulse, super wideband, time domain, carrierless, and other related concepts.

Before about 1989, UWB technology literature generally used all of the associated terms. Because UWB is a new term, it is best to look for the writer's definition or to determine the meaning in context and the accompanying details, and then apply the mathematical descriptions loosely. Same physical reason for assigning the breakpoints between narrowband, wideband, and UWB would be much more satisfying. Interpretation of systems as UWB should be kept loose. For example, an argument that a system with a 24 % fractional bandwidth signal is wideband and not UWB defies common sense and engineering judgment. Other examples of UWB usage include describing narrowband receivers and devices with a broad tuning range or broad proportional bandwidth RF amplifiers with a 1 to 18 GHz bandwidth.

1.1.4 Potential applications of UWB radar

Fine spatial resolution, extraction of target feature characteristics, and low probability of interception and non-interfering signal waveform are some of the features that make UWB radar appealing. Thus, UWB radar otters possible solutions to defense requirements such as passive target identification, target imaging and discrimination, and signal concealment from electronic warfare equipment and antiradiation missiles. Frequency spectrum sharing with other radar and communications systems is another potential use. Future UWB radar applications will depend on the ability of a particular UWB system to perform a given detection or remote sensing function competitively with available alternative systems or to provide same operational advantage, such as a low probability of intercept signal.

1.2 Objective of this work

Frequency-domain radar measurements have been dominant through the past and are usually performed using continuous or swept-frequency signals. In contrast, time-domain radar measurements are gaining more attention. A low-cost baseband pulsed microwave radar sensor was developed at our department based on our pulsed laser radar sensor technology. Extended-time sampling was used to downconvert the received picosecond pulses into the millisecond range, after which amplification, signal processing and digitizing of these downconverted pulses were performed [2]. Further software-based digital signal processing was used to display the target returns and to extract target range information from the received signals.

Fig. 1.1 presents the block diagram of the UWB radar in the bistatic configuration. The pulse sharpener circuit that feeds the radiating antenna uses a step recovery diode (SRD), which outputs pulses that are gaussian in shape, and have a full width half maximum (FWHM) of 150 ps and a rise time of 120 ps with a maximum pulse amplitude of 7 V. The frequency spectrum of each pulse extends from DC to 10 GHz with a bandwidth defined by the -20 dB (90% of pulse energy) of 5.95 GHz. The pulse repetition frequency (PRF) is adjustable from 20 kHz to 10 MHz.

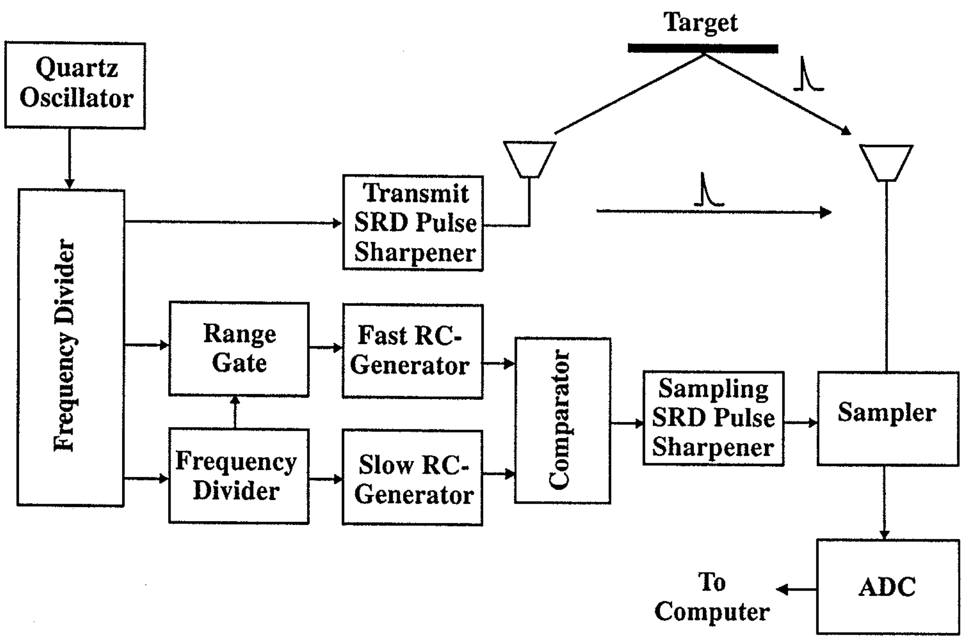


Fig. 1.1 A basic block diagram of the baseband pulsed microwave bistatic radar sensor [2].

There is also another pulse sharpener, which is also SRD-based, used to sharpen the sampling pulses. These pulses are also gaussian with FWHM of 70 ps and amplitude of 1.5 V, from which only the tip is used for sampling, making the sampling aperture less than 50 ps. The received pulses as well as the sampling pulses are fed into a DC-biased sampling Schottky diode bridge through 50 Ω microstrip transmission lines etched on a low-cost substrate.

Since extended time sampling is employed, the sampler output pulses that are in the millisecond range, which are fed directly into a high-gain transimpedance amplifier that also functions as an active low-pass filter with a -3 dB cut-off frequency of about 12 kHz.

The system uses two UWB antennas in a bistatic configuration for the transmission and reception of picoseconds pulses. Each antenna is a vertical inverted trapezoidal one and fed through an image ground plane and backed by a reflector [2]. The antennas were designed specifically for this system and have the capability to transmit and receive ultra short pulses with minimum distortion. Resistive loading was used in the antennas to reduce pulse reflections and multiple pulse transmissions. In the previous configuration the two antennas were spaced 15 cm apart, and amounted on a three legged stand.

To increase the mobility of the system and to avoid the undesired reflections from the system itself, the monostatic configuration is needed instead of the bistatic configuration, for this purpose signal divider and directional coupler implementation is the aim of this work.

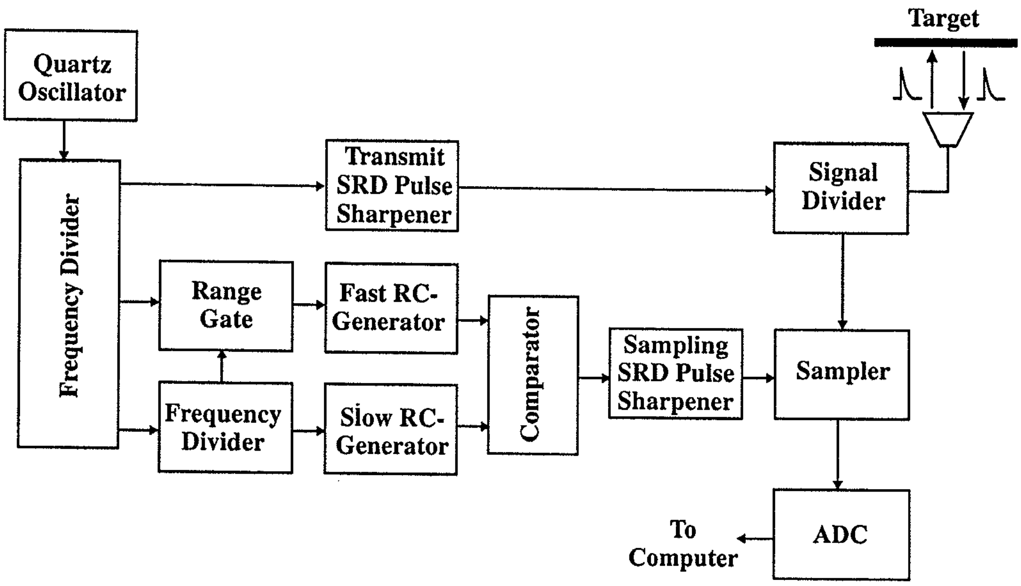


Fig. 1.2 A basic block diagram of the baseband pulsed microwave monostatic radar sensor.

Based on the realized UWB bistatic radar, a concept of an UWB monostatic radar has been developed. For the system realization, an UWB coupler was required for the detection of the transmitted and received signals. The device should be designed for ultra-short signals with a pulse width of about 100 ps.

The device was designed with ADS, fabricated and measured. A demonstrator of an UWB monostatic radar with the new device should then be built up including earlier developed components used for the bistatic radar system.

The difference between the two systems is shown in Fig.1.1 and Fig. 1.2. The main system components are the same in both systems. However, in the bistatic radar case there are two antennas, one for transmitting and the other one for receiving the reference signal as well as the reflected signal from the target. Both signals will be fed into the sampler. In the monostatic radar system only one antenna is used for transmitting and receiving UWB signals, with the aid of our coupler, which will couple those signals directly into the sampler.

1.3 Directional coupler and power divider

A directional coupler is a four-port microwave junction. And commonly comprises two straight-line waveguides which are coupled in someway. With reference to Fig. 1.3, which is a schematic illustration of a directional coupler, the ideal directional coupler has the property that a wave incident in port 1 couples power into ports 3 and 4 but not into port 2. Similarly, power incident in port 3 couples into ports 1 and 2 but not into port 4. Thus port 2 and 4 are uncoupled. For waves incident in port 1 or 3, the power is coupled into ports 4 and 2 only, so that ports 2 and 4 also uncoupled. In addition all four ports are matched. That is, if three ports are terminated in matched loads, the fourth port appears terminated in a matched load, and an incident wave in this port suffers no reflection.

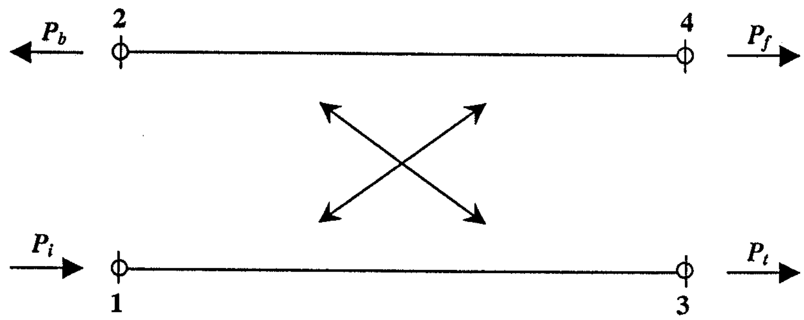


Fig. 1.3 The schematic illustration of a directional coupler.

Directional couplers are widely used in impedance bridges for microwave measurements and for power monitoring. For examine, if a radar transmitter is connected to port 1, the antenna to port 2, a microwave detector to port 3, and a matched load to port 4, the power received in port 3 is proportional to the power flowing from the transmitter to the antenna in the forward direction only. Since the reflected wave from the antenna, if it exists, is not coupled into port 3, the detector monitor the power output of the transmitter.

If the coupler is designed for 3 dB coupling, then it splits the input power in port 1 into equal powers in ports 3 and 4. Thus a 3 dB directional coupler serves as a power divider. Directional couplers with 3 dB coupling are also called hybrid junctions and are widely used in microwave mixers and as input and output couplers in balanced microwave amplifier circuits [2].

There are many available designs and configurations for directional couplers, hybrid junctions, resistive bridges, and power dividers will be discussed shortly in the next chapter.

S matrix of a forward coupler with port notation given in the device symbol:

$$S = \begin{bmatrix} \rho & \sigma & \tau & \kappa \\ \sigma & \rho & \kappa & \tau \\ \tau & \kappa & \rho & \sigma \\ \kappa & \tau & \sigma & \rho \end{bmatrix} \tag{1.3}$$

For abbreviation always two ports are joint to a port group:

$$S = \begin{bmatrix} \underline{S}_{11} & \underline{S}_{12} \\ \underline{S}_{21} & \underline{S}_{22} \end{bmatrix} \tag{1.4}$$

For reason of symmetry the 2-row submatrices have the properties:

$$\underline{S}_{11} = \underline{S}_{22}, \quad \underline{S}_{12} = \underline{S}_{21}$$

This means that only 4 scattering coefficients are required for characterization:

ρ reflection coefficient (eigen-reflection coefficient)

τ transmissions coefficient transmission path (main transmissions coefficient)

κ transmissions coefficient coupling path (coupling coefficient)

σ transmissions coefficient separated path

1.4 Directional coupler and power divider parameter definitions

The performance of the directional coupler is measured mainly by two parameters, the *coupling* and the *directivity*, let P_i be the incident power in port 1, P_f be the coupled power in the forward direction in arm 4 and P_t is the output power in arm 3.

Coupling

The power ratio in dB by which the coupled output port is *decoupled* from the input port when all ports are terminated in reflectionless (matched) terminations. The nominal coupling is specified as the arithmetic average of the maximum and the minimum coupling within the frequency band.
The coupling in decibels is then given by

$$\mathrm{C/dB} = 10 \log \left(\frac{P_i}{P_f}\right) = 20 \log \left(\frac{1}{|\kappa|}\right) \tag{1.5}$$

Directivity

Ideally, the power P_b coupled in the backward direction in arm 2 should be zero. The extent to which this is achieved is measured by the directivity D, which is defined as a power ratio in dB, so directivity is a measure of the preferential coupling of RF energy from the mainline to the coupled port in the "forward" direction compared to that in the "reverse" with all ports terminated in matched (reflectionless) loads. The directivity in decibels is then given by

$$\mathrm{D/dB} = 10 \log \left(\frac{P_f}{P_b}\right) = 20 \log \left|\frac{\kappa}{\sigma}\right| \tag{1.6}$$

Coupling tolerance

The specified allowable unit-to-unit variation in dB in nominal coupling. That means when the coupling tolerance of the device is so small, i.e. the nominal coupling variations are so small all over the frequency range.

Frequency sensitivity

The peak-to-peak deviation in dB from the nominal coupling over the specified frequency range, which is very important in UWB technology.

Insertion loss

The net unrecoverable power in dB dissipated within the circuit at any frequency within the specified range. The insertion loss is usually specified as *excluding* the coupling loss, but to avoid a complex and perhaps critical calculation, it is expressed as *including* coupling loss for dual directional and high power models. The insertion loss in decibels is then given by

$$\text{Insertion Loss/dB} = 10\ \log\left(\frac{P_t}{P_i}\right) \tag{1.7}$$

Coupling loss

The main line signal loss in dB attributable to power being sampled at the coupled port. The coupling loss in decibels is then given by

$$\text{Coupling Loss/dB} = 10\ \log\left(\frac{P_f}{P_i}\right) \tag{1.8}$$

Transmission loss

The sum of the maximum insertion loss and coupling ''loss'' in dB at any frequency in the specified range. The transmission loss in decibels is given by

$$\text{Transmission Loss/dB} = 10\ \log\left(\frac{P_t + P_f}{P_i}\right) \tag{1.9}$$

Ideal directional couplers:

With $\rho = 0$ and assuming that the device is lossless, it follows from the unitarity relation with $\tau, \kappa \neq 0$ and $\underline{S}_{11} = \underline{S}_{22} = 0$, so that the forbidden path is totally separated, i.e. $D = \alpha$. Furthermore, it turns out that $\underline{S}_{12} = \underline{S}_{21}$ unitary or

$$|\tau|^2 + |\kappa|^2 = 1, \qquad \tau^*\kappa + \kappa^*\tau = 0 \tag{1.10}$$

Both equations ca be fulfilled by the relation

$$|\tau| = \sqrt{1 - |\kappa|^2}\ , \quad arc(\tau) = arc(\kappa) \pm \frac{\pi}{2} \tag{1.11}$$

if the reference planes are suitably chosen then $arc(\tau)=0$ or $\tau = real$ and $|\kappa| = K$

$$S = \begin{bmatrix} 0 & 0 & \sqrt{1-|\kappa|^2} & \pm j\kappa \\ 0 & 0 & \pm j\kappa & \sqrt{1-|\kappa|^2} \\ \sqrt{1-|\kappa|^2} & \pm j\kappa & 0 & 0 \\ \pm j\kappa & \sqrt{1-|\kappa|^2} & 0 & 0 \end{bmatrix} \tag{1.12}$$

S matrix of an ideal 3 dB coupler ($\kappa = \frac{1}{\sqrt{2}}$ **)**

$$S = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 0 & 1 & \pm j \\ 0 & 0 & \pm j & 1 \\ 1 & \pm j & 0 & 0 \\ \pm j & 1 & 0 & 0 \end{bmatrix} \tag{1.13}$$

1.5 Directional coupler and power divider classes

A directional *coupler* may be used to sample either a forward or backward wave in a transmission line. A *unidirectional* coupler has available terminals or connections for sampling only one direction of transmission; a *bidirectional coupler* has available terminals for sampling both directions.

A directional coupler inserted in a transmission line allows precise monitoring of the RF energy flow in that line while introducing minimum perturbation of the main line signal in the sampling process.

Directional couplers are available in three different classes:

1) Three port directional couplers
2) Four port bi-directional couplers
3) Four port dual directional couplers

1.5.1 Three port directional couplers

Three port directional couplers are four port networks where one port (port 4) is internally terminated in a resistive load thus becoming the isolated port.

The three ports are:

1) Main line input port
2) Coupled output port
3) Main line output port

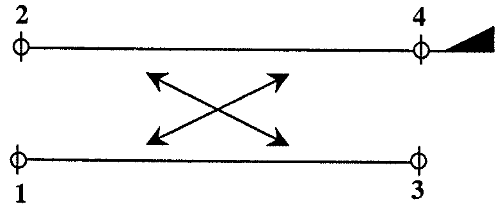


Fig. 1.4 *Three port directional coupler.*

An RF signal applied to the input port splits unequally between the coupled port and the main line output port. The degree of unequal power division is a function of the coupling ratio of the coupler.

The isolated port, commonly known as a "load port" is designed to absorb and dissipate reflected energy, stripline coupler designs are generally more appropriate suitable for direct integration into microstrip mother boards.

For transmitter and power amplifier systems requiring high power monitoring and leveling, directional couplers using an air dielectric are more suitable.

1.5.2 Four port bidirectional couplers and power dividers

The four port network has the isolated port externalized with either an RF connector or pin. One advantage of this type coupler is that a higher power termination can be selected to suit higher input power requirements. Ports 1 and 3 will be the input and output ports respectively, while ports 4 and 2 will be the forward and reverse coupled ports respectively.

The four port network can also act as a bi-directional coupler monitoring signals in both directions provided the coupled ports feed into a reasonably constant 50 Ω impedance.

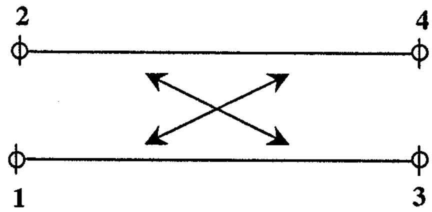


Fig. 1.5 *Four port bi-directional coupler.*

1.5.3 Dual directional couplers

Dual directional couplers (see Figure 1.6) are four port networks that are distinguished from the bi-directional types in that dual directional couplers are two independent four port networks connected in series. Ports 1 and 3 will be the input and output ports respectively, while ports 4 and 2 will be the forward and reverse coupled ports respectively. Ports 7 and 8 are terminated with matched loads, and ports 5 and 6 are directly connected to arise in the main line. Their principal application is in monitoring signals simultaneously in both directions.

They are preferred over the simple bi-directional coupler in that they provide higher isolation between coupled ports. Occasionally, a requirement arises that requires a dual directional coupler monitoring power in the same signal direction.

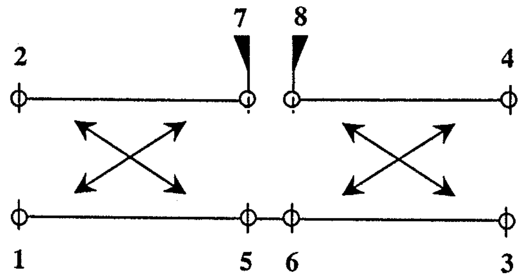


Fig. 1.6 Dual directional coupler.

These can be required for some special applications. Many microwave systems such as multi-channel receivers, transmitter and antenna systems require uniform phase and amplitude characteristics.

1.5.4 General comments about the directional couplers classes

A directional coupler is basically a 4-port network. The main-line and auxiliary line each have 2 ports: A 3-port coupler has one end of this auxiliary line, the "isolated port," internally terminated. When all 4 ports are made available to the user, the device is called a "bi-directional coupler."

A 3-port coupler have an advantage over a 4-port when only 3 ports are needed because directivity of a coupler is strongly affected by the impedance match provided by the termination at the isolated port. Furnishing that termination internally ensures high performance.

A 4-port coupler can be used to sample forward and reflected power simultaneously, by placing measuring instruments at both ports of the coupled line, but with care to provide good impedance match at all 4 ports of such a "bi-directional coupler." A coupler's directivity can be no better than the return loss of the terminations at the far-end main-line and coupled line ports; poor directivity causes inaccurate power monitoring by leaking forward and reflected signals into one another's paths. An alternative approach which overcomes this limitation is to use two 3-port couplers back-to-back; this combination is called a "dual directional coupler."

Directivity is measured in the case of the three port directional coupler by measuring the the loss from the main-line input to the coupled port with the main-line output terminated. Then reverse the main-line connections. The difference in dB readings is the directivity.

A coupled port can be either an input or an output, The coupling factor determines the attenuation between main-line and coupled line signals in both cases.

All ground pins should be connected to ground with short path length to obtain full directivity and VSWR performance.

1.6 Directional couplers and power dividers applications

1. Directional couplers are used to accurately sample the directional power flow in a transmission line. In conjunction with a calibrated detector or a power bridge, an accurate, continuous measurement of power flow can be obtained. In this function directional couplers can be an essential part of system BITE (built in test equipment).

2. Power leveling can be performed when the coupled output of a directional coupler is used in conjunction with a modulator or a PIN attenuator as a part of a leveling loop.

3. Frequency measurement can be made on a continuous basis when the coupled sample is fed to a suitable frequency counter or equivalent.

4. Frequency stabilization can be obtained when the coupler output is used as the input to an AFC (automatic frequency control) loop.

5. Continuous power reflection measurements such as might result from antenna misalignments can be made using couplers as reflectometers. The power source is fed to the main line *output* port and the coupler main line *input* port is connected to the load (antenna). Reflected power will thus be coupled to the *forward* coupled output where it can be monitored.

6. Signal injection can be obtained by feeding the injectable signal into the mainline via the coupled port. The direction of the inserted signal power flow depends on the coupler polarity in the transmission line.

7. Distribution of a low power signals to two or more antennas.

Chapter 2 :

Baseband pulsed microwave radar sensor

2.1 Introduction to pulsed radar sensor

2.2 Developed UWB baseband microwave radar

Chapter 2: Baseband pulsed microwave radar sensor

2.1 Introduction to pulsed radar system

Principle purpose of the pulse radar system is to search, detect and measure the range of targets. This system is the all rounder of radar system and with the addition of special sub-systems can perform other roles. Doppler radar is an example of an extension of the basic pulsed radar system.

As the name implies this radar system sends out a periodic pulse. This is not quite the case as a carrier is used as and so a group of pulses is sent out, i.e. a pure carrier is transmitted for the pulse period.

2.1.1 Operation of the pulsed radar system

The width of the pulse can vary between picoseconds and nanoseconds. In our UWB radar it is 150 ps. The pulse width determines the ability of the radar to distinguish between targets, and also determines the radar resolution. The transmitter unit transmits the UWB pulse, then waits for the echo.

If the echo is received ΔT seconds later then the range can be determined as:

$$R = \frac{C \times \Delta T}{2} \tag{2.1}$$

Where R is the distance between the target and the radar, and C is the velocity of the light which is equal to 3×10^{8} m/s. The transmitter does not wait indefinitely for the echo as there is a maximum range from which a targets echo is so weak it can not be detected. Therefore the transmitter waits for inter pulse period (IPP) which dictates the maximum range, R_{max}, which the pulsed radar system can detect a target. The inverse of the IPP is the PRF.

To measure the time delay it takes for the echo to reach the receiver we need a reference point in the transmitted signal. The echo that will be picked up by the receiver from the target will be an attenuated version of the transmitted signal and so its shape will be very similar to that of the transmitted pulse. The pulse shape to be transmitted therefore needs to have, one and only one, sharp reference point.

2.1.2 Range ambiguity:

Range ambiguity results from the fact that we only wait a limited period of time for an echo from a target before the next pulse is transmitted. Range ambiguity occurs when if for some reason we get an echo from a distance greater then R_{max}, i.e. after a second pulse has been transmitted. The receiver then can not tell from what range the echo came from.

For instance, if the target echo was detected 5 μs after a pulse, and the IPP is 0.6 ms. R_{max} for this system will be calculated as $R_{max} = IPP \times C/2 =$ 90 km. The echo could therefore have come from a range R of 750 m or 90.75 km according to equation 2.1.

It is therefore the IPP or the PRF that determines the amount of range ambiguity as shown in Fig. 2.1. The range in this figure could be R1 or R1+R2.

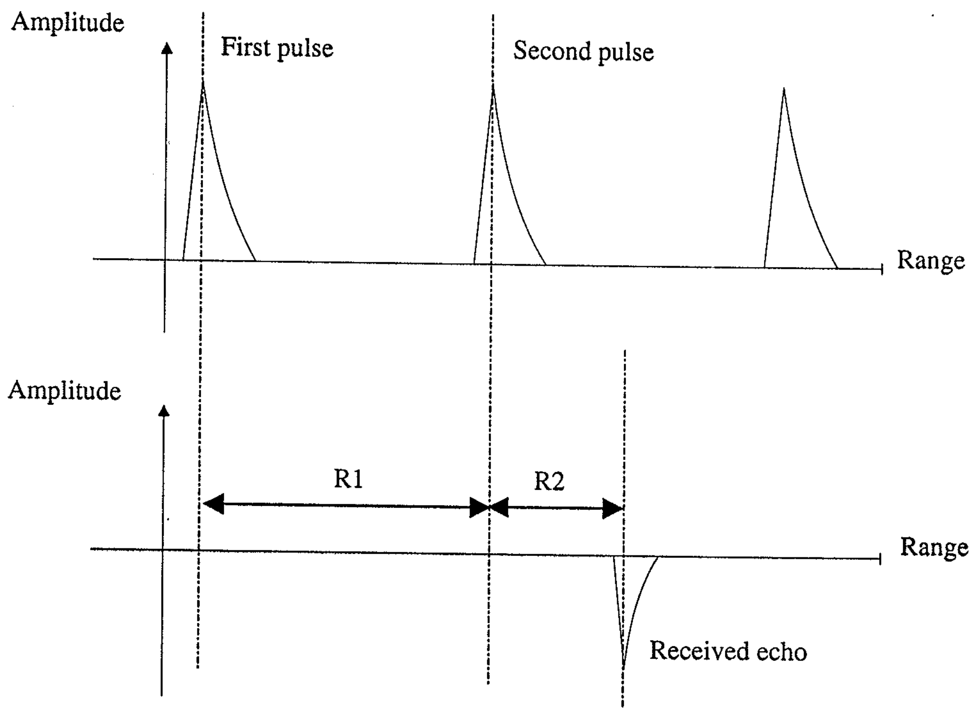


Fig. 2.1 Determination of the amount of range ambiguity.

If we set the PRF to a large enough value we can be certain we will not get any echoes from greater then R_{max}. But there are other factors like antenna rotational speed that limit the PRF value.

2.1.3 Range Resolution:

Range resolution is the ability of the system to distinguish between two targets that are closely positioned.

The echoes of the two targets must therefore not overlap to such an extent that they can not be still recognized as two separated echoes. Therefore, the shorter the pulse duration period the higher the range resolution, that is why we have used 150 ps UWB pulse to have a very high resolution.

2.1.4 The definition of the characteristics of a pulse sequence:

The definition of the characteristics of a pulse sequence is shown in Figure 2.2

Where T_r is the rise time,
T_f is the fall time,
T_i is the pulse width,
T is the period,
A is the amplitude and
a is the overshoot.

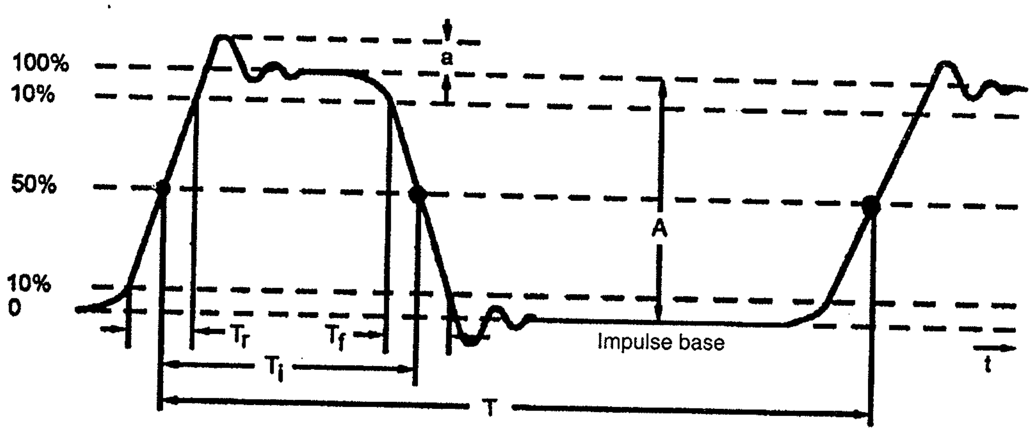


Fig. 2.2 Definition of the characteristics of a pulse sequence.

2.2 Developed UWB baseband microwave radar

To satisfy the requirement of a monostatic microwave radar sensor for near range measurement, a series of modules must be used. All of these modules are already used in the bistatic radar system, the only added module to the system is the "UWB directional coupler and signal divider" as it is already described before.

These modules include:

Pulse generator, ΔT circuit which consists of fast RC circuit and slow RC circuit, pulse sharpening circuits, sampler, antennas, amplifier and A/D converter. The system was shown in Fig. 1.2.

A quartz stabilized oscillator is used in this system to oscillate and to generate a signal of 20 MHz. Using a digital frequency divider circuit, the frequency of this 20 MHz signal is downconverted to 40 kHz. Then, this signal is fed into the low and fast RC circuits, SRD pulse sharpening circuits, sending antenna, sampler and finally to the computer using A/D converter.

Sequential sampling technique is used in this system as will be described in the next section. Slow and fast RC circuits form the ΔT circuit, which shifts the pulses after each sampling event by a small time interval ΔT. An avalanche transistor circuit, as a first stage, sharpens the rise time of the input signal to about 50 ns, and also amplifies its amplitude to about 5-15 Volts.

After that, SRD sharpening circuit is used as a second sharpening stage to sharpen this 50 ns pulses and to generate Gaussian pulses with rise time of about 130 ps and peak voltage of 5.5 V. These ultra-short pulses are fed through a 50 ohm coaxial cable into port 1 of the directional coupler, which couples it to the antenna at port 2 and to the sampler at port 3 as a reference pulse.

The reflected signal from the target will be received by the antenna and it will be fed into port 2 of the directional coupler, to be coupled into port 3 to the sampler as an echo. The sampler samples the two signals, the reference and the echo.

The sampler is controlled by the sampling pulses generated from another SRD pulse sharpening circuit. The rise time of this sampling pulses is 75 ps and amplitude of about 1.6 V.

2.2.1 Ultra-short electrical pulse generator

Ultra-short Electrical Pulse Generator that can deliver pulses with amplitudes of few volts and rise-time of 130 ps has applications, in signal transmission and sampling processes.

2.2.1.1 Pulse generator

In UWB monostatic radar system ultra-short pulses are transmitted from the antenna, then reflected back from the target and received again by the antenna.

In addition, the received pulses have to be sampled using other pulses with faster rise times and narrower widths. The pulse sharpening circuits based on SRD have to be fed by a first stage generator circuit. An example of a pulse generator is this one used in the system which consists of a 20 MHz-oscillator, frequency divider and a ΔT-circuit that shifts the sampling pulses.

The oscillator generates a train of rectangular pulses with 50% duty cycle. The rectangular pulses have a width of $1 / (2 \times PRF) = 25\ ns$ and a transition time of about 5 ns.

Finally, an SRD pulse sharpener circuits are used to sharpen these pulses and to obtain Gaussian picosecond pulses which can be used for transmitting and sampling [3].

2.2.1.2 SRD pulse sharpener circuit

The SRD can be utilized for many applications. The main application area lies in the field of pulse sharpening and signal shaping. The SRD acts like a switch whose properties depend on the instantaneously stored charges. The diode represents a low impedance under forward operation, when changes have flown into the diode. If the charge is withdrawn by a negative voltage, then the diode remains in the low-impedance situation for the first time as long as the charges have not been completely removed. However, after total remove the diode switches abruptly from the low impedance to a high impedance value.

This capability of the SRD, to store charges, and to change its impedance very fast is suited to generate impulses with sharp leading and falling edges, and for pulse shaping as well. Under pulsed operation the way in which the SRD changes its impedance value critically depends on the circuit design. The SRD principle of operation is described clearly in Fig.2.3.

The SRD is similar to the construction of a PIN diode with its static (DC) characteristics of a common p-n diode. Beyond this their dynamic (switching) behavior are quite different [3].

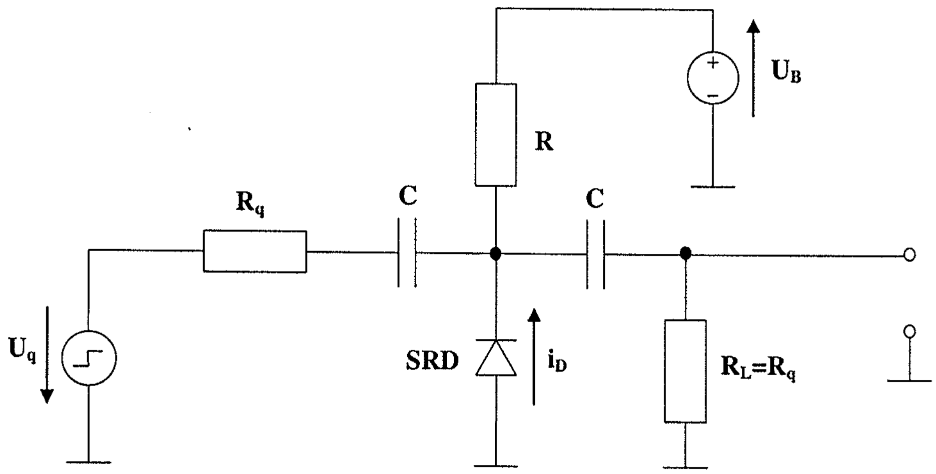


Fig. 2.3 Step recovery diode circuit configuration [3].

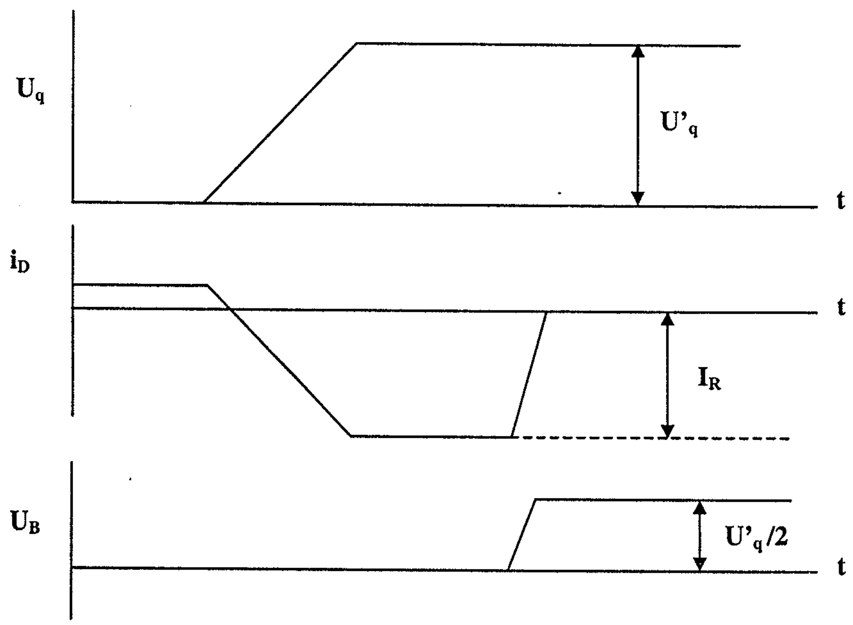


Fig. 2.4 SRD characteristics and principle of operation [3].

2.2.2 Sampler

For high resolution performance of measuring instruments, fast pulses are needed. Generally, the accurate determination of the significant pulse parameters may become very difficult when operating with extremely short pulse shapes having broad spectral content. The sampling technique, which is widely used today in analog signal processing and data conversion systems, is well studied for spectrum translation, as it makes simpler low frequency measurement techniques applicable to obtain the properties of actual high frequency signals [3].

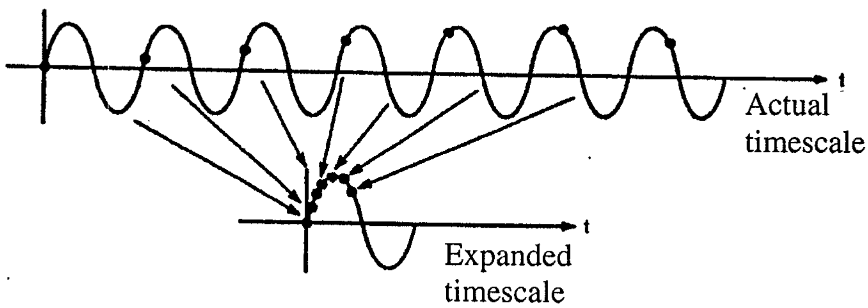


Fig. 2.6 The principle of the sequential sampling techniques[3].

So, in our system it is clear that the application of the sampling technique is very useful when broadband pulses (150 ps) are to be recorded and amplified. It is much easier to extent such a signal by sequential sampling, then to deal with this lower frequency sampled signal, than dealing directly with this UWB signal. The electrical expenditure is relatively low. High sensitivity and high resolution capability is involved. Noise is the limiting factor of the system sensitivity.

Fig. 2.6 illustrates the sampling technique which can be considered as an electronic version of the well known stroboscope, and was first applied to cathode-beam oscillographs. There are two main types of sampling; sequential sampling and random sampling. In the case of the sequential sampling, the sampling gate pulse is shifted after each sampling cycle by a suitable small time interval ΔT. That is why ΔT circuit is needed in the system; While in the random sampling case, this time intervals ΔT are not constant but follow in a statistical sequence. This random sampling method can also used for measuring very low pulse repetition rates, even if the longer observation time, and the relative higher statistical scattering may be regarded as disadvantages.

2.2.2.1 *Sequential sampling*

The operation of the sequential sampling is illustrated in Fig. 2.7. The measuring signal reaches the sampling gate via the trigger junction circuit and a delay line.

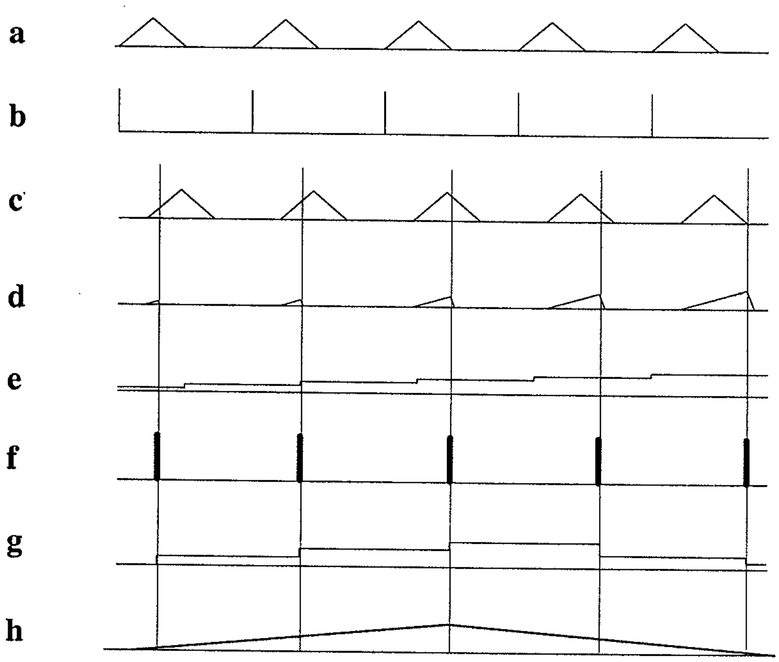


Fig. 2.7 *Time-dependent signals for sequential sampling.*
(a) measuring signal, (b) trigger pulse, (c) delayed signal
(d) saw tooth, (e) step voltage, (f) sampling gate pulse
(g) stored vertical voltage, (h) recorded signal.

The delay time must be equal or larger the response time of the saw tooth generator. Part of the undelayed signal is coupled out of the trigger junction circuit. This part must be carefully designed to prevent any distortion of the measuring signal The trigger impulse is shaped and starts the saw tooth generator (fast RC circuit).The fast saw tooth determines the regarded real time region.

Simultaneously with each cycle the step generator is stimulated (could be also slow RC circuit) in such a way that with each trigger impulse the step voltage is increased by one defined step value. Both signals, saw tooth and step (fast RC and slow RC) are compared in comparator circuit, which always generates a short impulse at the output when the actual values of the saw tooth and the step are the same. Now the sampling gate pulse opens the sampling gate for a short time, and the time-value of the signal to be measured is stored in vertical storage. The corresponding voltage value is amplified by an amplifier with relatively high gain.

The sampling cycle is determined by the repetition frequency of the signal to be measured. It is always assumed that the input signal does not change its waveform.

2.2.2.2 *Sequential sampling system*

The sequential sampling system is illustrated in Fig. 2.8. The output signal of the frequency divider controls a fast RC generator with a time constant T_f = 100 ns and a second frequency divider which converts the frequency to 20 Hz.

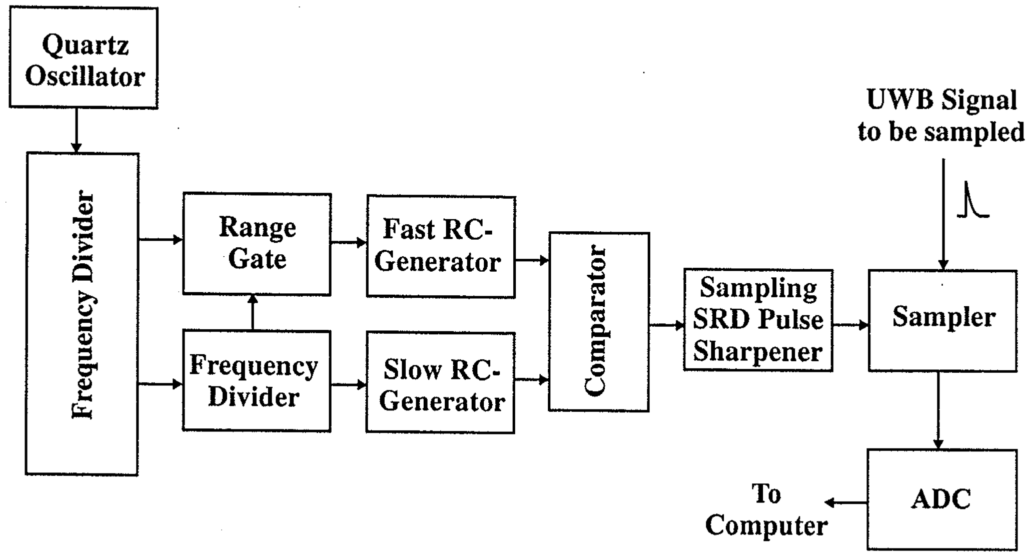


Fig. 2.8 A basic block diagram of the sequential sampling system.

This second frequency divider controls a slow RC generator with a time constant T_s = 375 ms. The output voltages of the two RC generators are compared by a comparator which produces a pulse when the voltages are equal. Then the pulse is fed into the SRD sharpening circuit and used for sampling.

Since the controlling signal of the sampling pulses comes from the ΔT generator, where each clock signal is shifted by a fixed shift ΔT of 6 ps, a sequential sampling technique known as extended-time sampling is realized. This operation causes the sample output to be extended by a factor of 10^6 in time domain, or to be downcoverted in the frequency domain from the GHz to the kHz rang. Therefore, using low frequency, "off the shelf" components such as operational amplifiers will be possible. This in turn helps to reduce the number of high-frequency components needed, which reduces the overall cost of the sensor. On the other hand this kHz range signal could be converted to a digital signal using A/D converter, then we can apply signal processing techniques.

2.2.2.3 Sampling gate

A symmetric construction of the sampling gate is shown in Fig. 2.9. This circuit provides the required decoupling. Because of the voltage drop over the resistors, the sampling efficiency is rather law. The diodes are again biased in backward direction, and are opened by the balanced gate pulses , These balanced gate pulses cancels the effect of each other on the input signal [2].

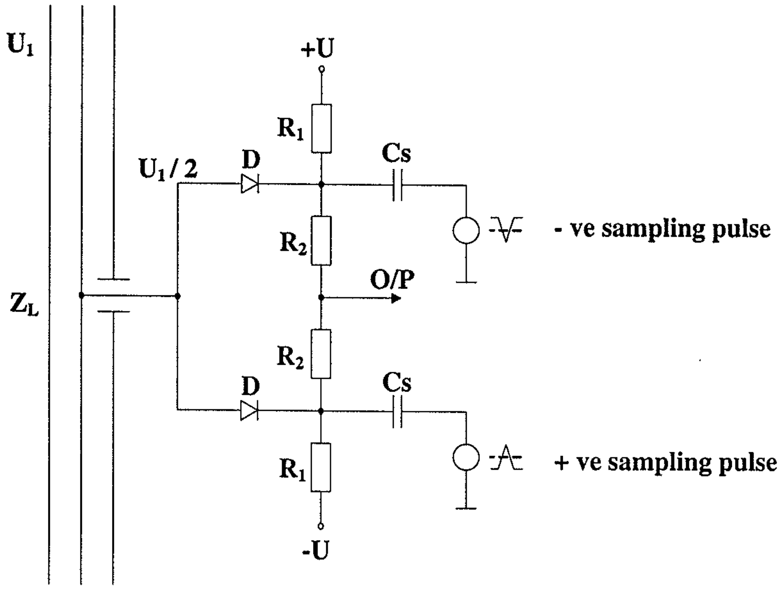


***Fig. 2.9** Simple sampling gate in form of a semi-bridge circuit [3].*

The most important advantage of the balanced sampler is that it approximately cancels the "kick out" effect and may be realized in a floating earth, DC coupled, configuration as per the two-diode sampling gate.

Kick out is defined as the leakage of the sampling pulse from the sampler to the input signal. A balanced circuit can be used to suppress the kick out by about 35 dB. This is very necessary to prevent interference to and reflections from, the circuit under test, which is the signal divider in the monostatic radar system.

In this circuit the signal is sampled by balanced strobe pulses applied through capacitive coupling. Because the input line is floating and the signal may be amplitude asymmetric, the diodes can be expected to assume an unequal charge store state. An output is thus related to both the amplitude and polarity of the input signal.

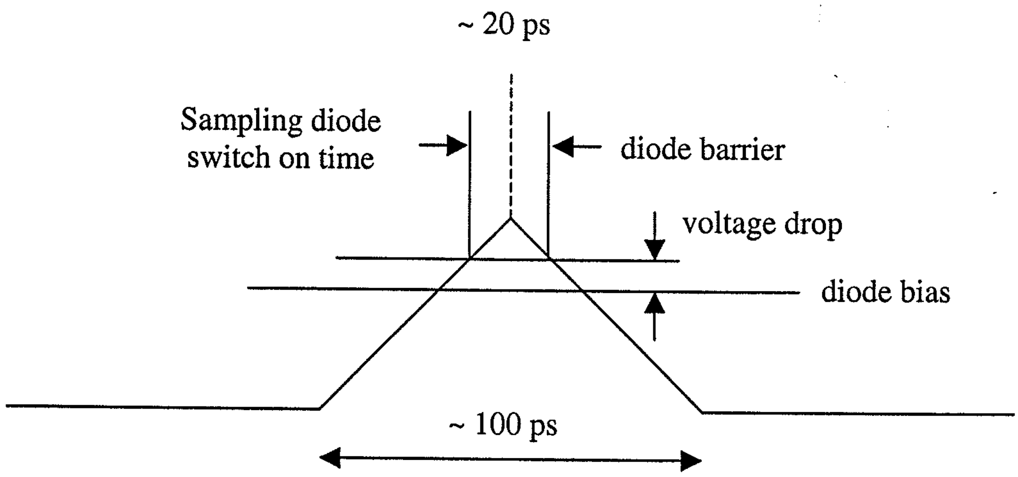


Fig. 2.11 Sampling diode bias [3].

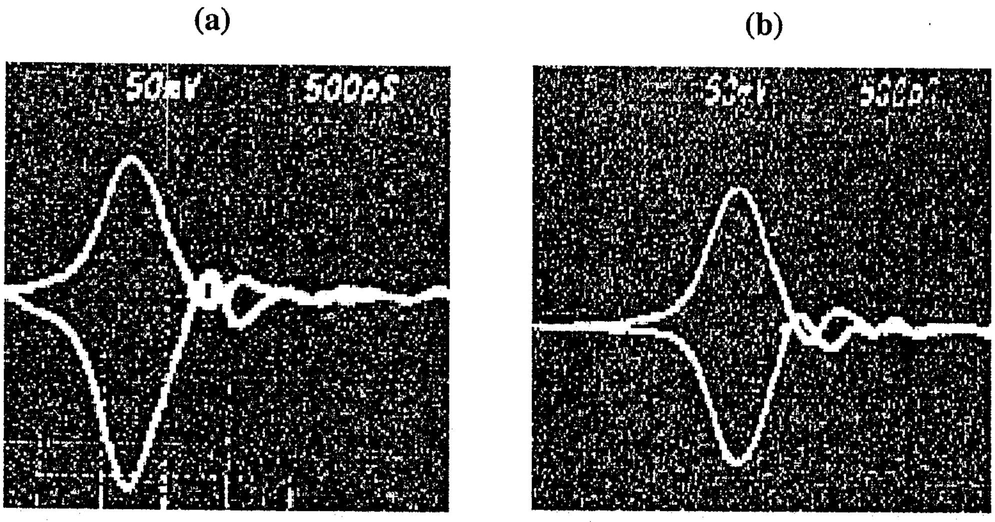


Fig. 2.12 (a) Measured balanced pulses without balance tuning.
(b) Measured balanced pulses after balance tuning [3].

A feedback path is provided to reset the reverse bias applied to the diodes after each sampling action so that subsequent samples only detect voltage changes in the input signal. This arrangement increases the dynamic range of the sampler and maintains a nearly optimum bias. Fig. 2.11 shows the principle of sampling diode bias.

2.2.3 UWB Directional Coupler and Signal Divider

The UWB directional coupler and signal divider is mainly a three port device. Its signal flow graph can be shown in Fig 2.13.

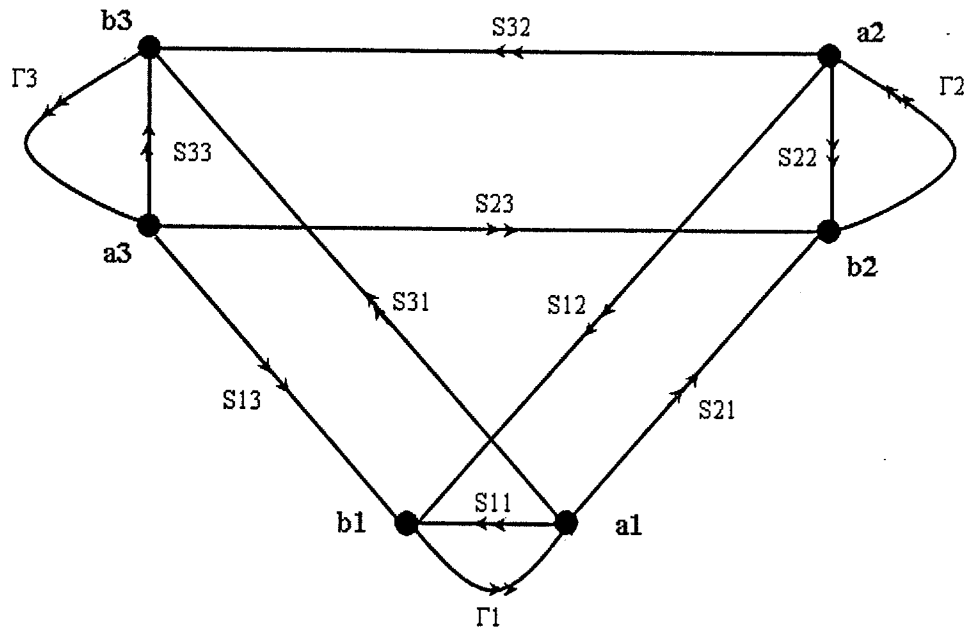


Fig. 2.13 Signal flow graph of three port directional coupler

Where :

S_{11}, S_{22}, S_{33} are the eigen reflection coefficients at ports 1, 2 and 3 respectively.
S_{12} is the transfer coefficients from port 2 to port 1.
S_{21} is the transfer coefficients from port 1 to port 2.
S_{13} is the transfer coefficients from port 3 to port 1.
S_{31} is the transfer coefficients from port 1 to port 3.
S_{23} is the transfer coefficients from port 3 to port 2.
S_{32} is the transfer coefficients from port 2 to port 3.
Γ_1, Γ_2, Γ_3 are the reflection coefficients at ports 1, 2 and 3 from the pulse generator, the antenna, and the sampler respectively.

Applying Mason's rule on this diagram [4] we can easily determine the transmission coefficients between the ports and also the reflection coefficients at each one. In our case for design and optimization, we will terminate the three ports with 50 Ohm termination. Therefore, we can consider that all of the three ports are fully matched and each port is terminated with a 50 Ohm load.
That means: $\Gamma_1 = \Gamma_2 = \Gamma_3 = 0$.

This assumption results in more simplicity to calculate the reflection coefficient at each port, which will be equal to the eigen reflection coefficient at this port. Also the transmission coefficient between any two ports will be equal to the transmission S-parameters coefficients between these two ports.

Our design of this three port directional coupler and signal divider mainly depends on very important two goals. The first one is to reduce the eigen reflection coefficients of this device as possible. i.e.: S_{11}, S_{22}, S_{33}. The second goal is related to the transmission coefficients S_{21}, S_{31}, S_{32}, which we need to increase their values as we can. All of these parameters can be shown in the signal flow graph presented in Fig. 2.13.

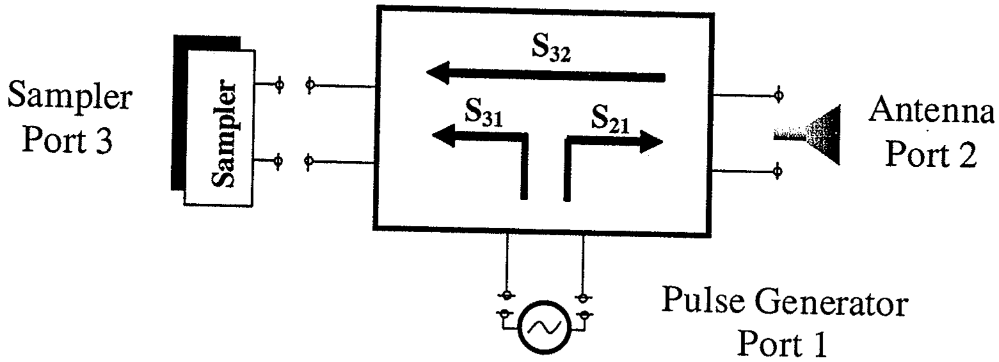


Fig. 2.14 Three port UWB directional coupler in our system.

The most important eigen reflection coefficients is S_{22}, because port 2 is directly connected to the antenna, therefore the reflected signal with small power shouldn't suffer from high reflection. In the mean time, it is easily concluded that the most important transmission coefficient, which should be maximized as we can is S_{32}, because this transmission goal S_{32} is the most important factor responsible about the low power reflected signal transmission from port 2 to port 3 as can be shown in Fig. 2.14.

Therefore, It is confirmed that it is so important to determine our signal divider goals as we have mentioned, and to take in our consideration that those goals should be frequency independent. i.e. The S-parameters of this signal divider should be approximately constant all over the frequency range D.C. - 20 GHz. If not, the transmitted and reflected UWB pulses will suffer from dispersion. That leads us to the basic idea of our design, which should mainly consist of lumped resistors.

This idea was mentioned actually in such other applications in reflectometer design and described as a resistive bridge [4] and it will be described in details in the following chapters.

Chapter 3 :

UWB directional coupler and signal divider design basis

3 UWB directional coupler and signal divider design basis

3.1 Power dividers and 3-port network

Power dividers are passive microwave components used for power division or power combining [5]. Power dividers can be three port networks in the form of T-junctions, and could be four port networks in the form of directional couplers and magic Tees. On the other hand, power dividers could be lossless components or lossy ones. Directional couplers can be designed for arbitrary power division, while hybrid junctions usually have equal power division. Hybrid junctions have either a 90° (quadrature) or a 180° (magic-T) phase shift between the output ports.

Power dividers are often of the 3 dB type, but unequal power division ratios are also possible. In this work, the case of power division is needed. Therefore, an input signal is divided by the three port coupler into two signals of lesser power. The unequal power division ratios in this work are required.

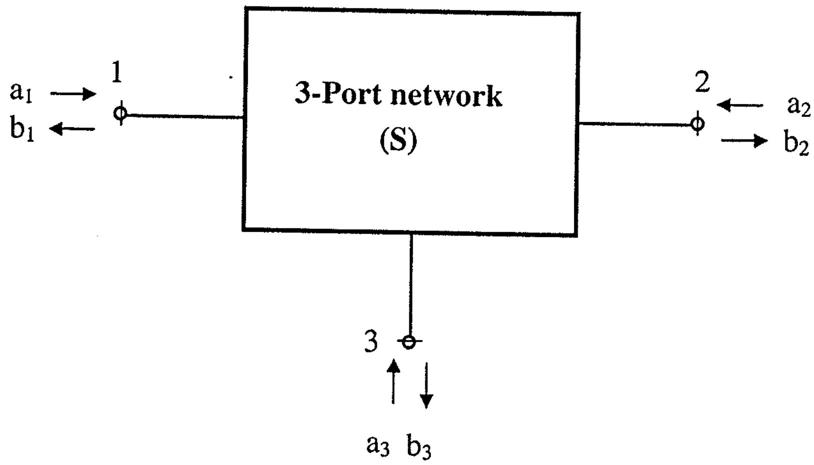


Fig. 3.1 Three port network power divider.

The simplest type of power divider is a T-junction.
The scattering matrix of a three-port network has nine independent elements:

$$S = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix} \tag{3.1}$$

If the components is passive and contains no anisotropic materials, then it must be reciprocal and its [S] matrix must be symmetric ($S_{ij} = S_{ji}$).

To avoid power loss, we would like to have a junction that is lossless and matched at all ports; however it is impossible to construct such a three-port lossless reciprocal network that is matched at all ports.

If all ports are matched, then $S_{ii} = 0$. If the network is also reciprocal the S matrix can be reduced to

$$S = \begin{bmatrix} 0 & S_{12} & S_{13} \\ S_{12} & 0 & S_{23} \\ S_{13} & S_{23} & 0 \end{bmatrix} \tag{3.2}$$

If the network is also lossless, then energy conservation requires that the scattering matrix be unitary, $[S]^t[S]^* = [U]$, which leads to:

$$\text{Condition 1}: \begin{cases} S_{13}^* S_{23} = 0 \\ S_{23}^* S_{12} = 0 \\ S_{12}^* S_{13} = 0 \end{cases} \qquad \text{Condition 2}: \begin{cases} |S_{12}|^2 + |S_{13}|^2 = 1 \\ |S_{12}|^2 + |S_{23}|^2 = 1 \\ |S_{13}|^2 + |S_{23}|^2 = 1 \end{cases}$$

Condition 1 implies that two of (S_{12}, S_{13}, S_{23}) must be zero, this condition will always be inconsistent with condition 2. If any of these three conditions is relaxed, then a physically realizable device is possible.

If the three-port network is non-reciprocal, then $S_{ij} \neq S_{ji}$ and the conditions of input matching at all ports and energy conservation can be satisfied. Such device is known as a circulator, and generally relies on an anisotropic material, such as ferrite, to achieve nonreciprocal behavior. Any matched lossless three-port network must be nonreciprocal and, thus, a circulator.

The [S] matrix of a matched three-port network has the following form:

$$S = \begin{bmatrix} 0 & S_{12} & S_{13} \\ S_{21} & 0 & S_{23} \\ S_{31} & S_{32} & 0 \end{bmatrix} \tag{3.3}$$

If the network is also lossless, then energy conservation requires that the scattering matrix be unitary, $[S]^t [S]^* = [U]$, which leads to:

$$\text{Condition 1 : } \begin{cases} S_{31}^* \, S_{32} = 0 \\ S_{21}^* \, S_{23} = 0 \\ S_{12}^* \, S_{13} = 0 \end{cases} \qquad \text{Condition 2 : } \begin{cases} |S_{12}|^2 + |S_{13}|^2 = 1 \\ |S_{21}|^2 + |S_{23}|^2 = 1 \\ |S_{31}|^2 + |S_{32}|^2 = 1 \end{cases}$$

Condition 1 and 2 can be satisfied in one of the following ways:

$$S_{12} = S_{23} = S_{31} = 0, \qquad |S_{21}| = |S_{32}| = |S_{13}| = 1$$

or

$$S_{21} = S_{32} = S_{13} = 0, \qquad |S_{12}| = |S_{23}| = |S_{31}| = 1$$

The result shows that $S_{ij} \neq S_{ji}$ for $i \neq j$, which implies that the device must be nonreciprocal. The [S] matrices for the above solution are:

$$S = \begin{bmatrix} 0 & 0 & 1 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \quad \text{or} \quad S = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 1 & 1 & 0 \end{bmatrix} \tag{3.4}$$

There is also another possibility for the three port network, It could be matched, reciprocal network, but it will be lossy in this case to satisfy the conservation conditions. One example of these networks is the resistive divider.

In this work, the *resistive divider* is preferred for the UWB monostatic radar system, because it gives better performance in controlling the coupling goals, also it could be designed to be approximately matched at all ports.

3.2 Resistive divider

If a three-port divider contains lossy components (could be lumpėd components) then it can be made to be matched at all ports. An equal-split three-port resistive power divider can be shown in the Fig. 3.2.

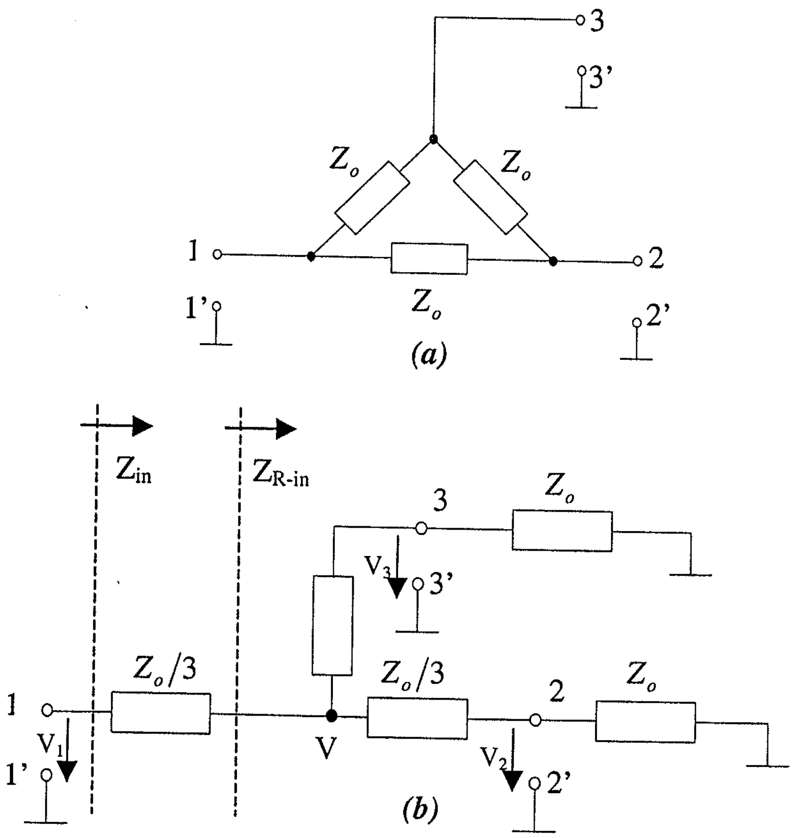


Fig. 3.2 An equal-split three port resistive power divider configuration.
(a) Delta configuration circuit (b) Circuit analysis

To calculate the input impedance of the circuit, it is considered that all of the ports are terminated with matched loads, also we have to use the delta-star conversion rule.

$$Z_{R\text{-}in} = (4Z_o/3) \,//\, (4Z_o/3) = 2Z_o/3 \tag{3.5}$$

$$Z_{in} = 2Z_o/3 + Z_o/3 = Z_o \tag{3.6}$$

From symmetry, that leads to $S_{11} = S_{22} = S_{33} = \dfrac{Z_{in} - Z_o}{Z_{in} + Z_o} = 0$

$$V = \frac{V_1}{Z_o} \times \left(\frac{2 \times Z_o}{3} \right) = \frac{2}{3} V_1 \tag{3.7}$$

$$V_2 = V_3 = \frac{V}{Z_o + (Z_o / 3)} \times Z_o = \frac{3}{4} V = \frac{V_1}{2} \tag{3.8}$$

From symmetry, it is clear that $S_{21} = S_{31} = S_{23} = ½$

Since the network is reciprocal, the S matrix can be written as:

$$S = \frac{1}{2} \begin{bmatrix} 0 & 1 & 1 \\ 1 & 0 & 1 \\ 1 & 1 & 0 \end{bmatrix} \tag{3.9}$$

The power delivered to the input of the divider P_{in} is

$$P_{in} = \frac{1}{2} \frac{V_1^2}{Z_o} \tag{3.10}$$

P_2 and P_3 are the output powers at port 2 and 3 respectively, can be estimated as

$$P_2 = P_3 = \frac{1}{2} \left(\frac{V_1}{2} \right)^2 \times \frac{1}{Z_o} = \frac{V_1^2}{8 Z_o} = \frac{1}{4} P_{in} \tag{3.11}$$

$$P_{out} = P_2 + P_3 = \frac{P_1}{2} \tag{3.12}$$

A simple analysis of this circuit will demonstrate that any one of the three ports has a 50 ohm input impedance when the other two are terminated in 50 ohms, and that the insertion loss between any two ports is 6 dB. A microwave network of this type consists of a symmetrical resistive film deposited on a ceramic substrate having three conducting contacts, each connected to the center conductor of a coaxial connector.

Resistive dividers provide well-matched signals of essentially equal magnitude and phase over a very broad band as opposed to the reactive and hybrid types which employ frequency limitive techniques. The resistive divider is intended for applications where the output signals are used independently, such as the simultaneous monitoring of power and frequency [6].

The previous three-port divider has the same transfer S-Parameters because of the symmetry of the circuit. When we have other different transfer goals, we have to redesign the circuit, and to choose the suitable values of the components to optimize our circuit to realize our required goals.

The resistive bridge used in the reflectometer configuration based upon delta configuration. This circuit was optimized to suit our application, also the star configuration of this circuit has been constructed using the delta to star conversion rule to reduce the effect of the parasitics of the moicrostrip discontinuities. Finally another circuit based upon the idea of the resistive potential divider was designed.

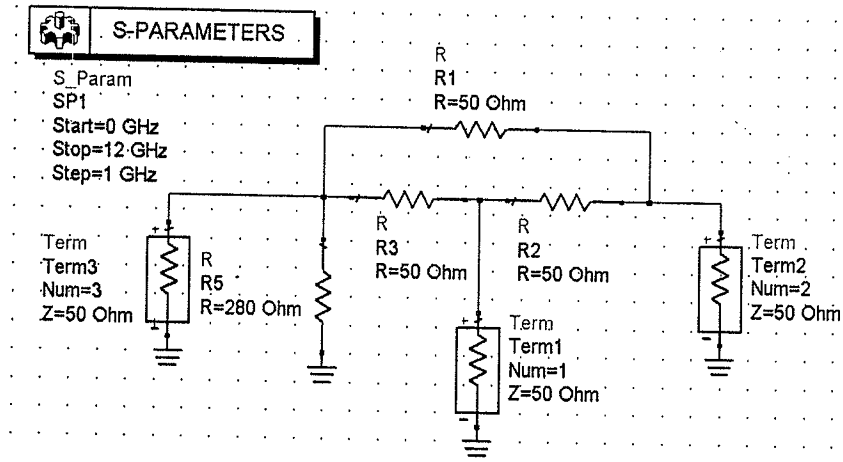
The optimization goals of the three circuits in the S_parameters simulation using the ADS are the same and they are summarized as :

Reflection goals	: S_{11}, S_{22}, S_{33}	should be minimum (-20 dB)
Transmission Goals	: S_{21}, S_{31},	should be accepted (-10 dB)
The most important goal	: S_{32}	should be maximum (0 dB)

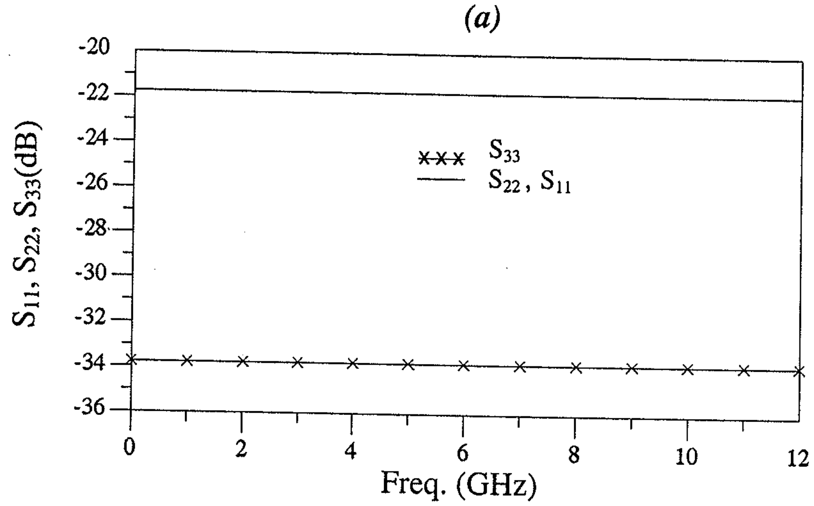
The well-known resistive bridge concept is usually used in the UWB range applications from a few MHz to at least 40 GHz in coaxial scalar network analyzers [4,5].

The resistive bridge is intended e.g. for a realization in a planar circuit (microstrip or MMIC technology) with coaxial connectors. In this case, the imperfect internal load of the bridge and the reflections at the waveguide transition to the coaxial system limit the directivity of the bridge. The directivity of the bridge only depends on the symmetry of the lines, the waveguide transitions and the bridge itself.

In Fig. 3.3 we have used only the Schematic simulation for the resistive bridge circuit as shown, and we didn't take the layout discontinuities effect in our consideration i.e. Tee junctions, crosses, angles, also it is considered that the components are ideal resistors, not the realistic used ones which includes parasitics. R_5 is added only for better stabilization. In the following sections in this chapter we will discuss all of these effects.



(a)



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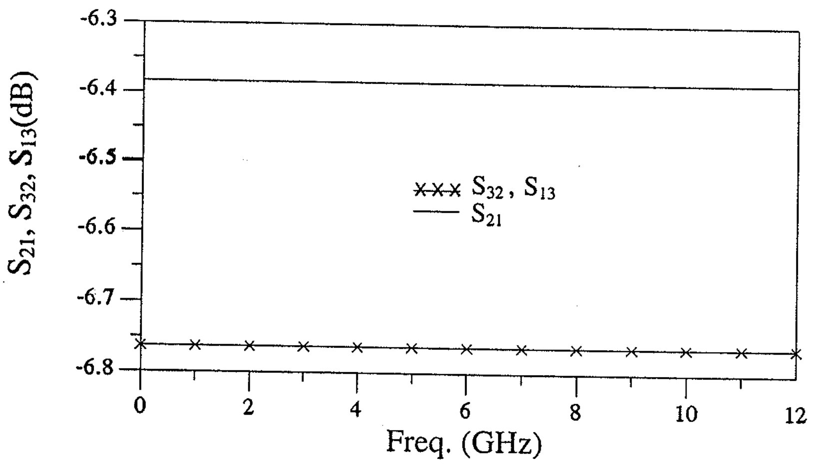

(c)

Fig. 3.3 *Resistive bridge S-parameters schematic simulation.*
(a) ADS schematic circuit. (b) Eigen Reflection coefficients.
(c) Transmission coefficients.

3.3 Microstrip discontinuities

Generally a discontinuity is any change in the geometry of a waveguide cross-section along the line. It disturbs the propagation of the electromagnetic, and leads to undesired effects such like reflection, attenuation and radiation of electromagnetic energy regarding open transmission structures.

On the other side they render possible the realisation of relevant devices commonly used in microwave and millimetre wave circuits, e.g. junctions, filters, matching elements and mode transformers. Also any change in the direction of waveguide longitudinal axis and the transition to any circuit element (bonded hybrid elements) as well, represents a discontinuity. The quantification is of great practical importance for circuit design.

It is known that the theoretical treatment of waveguide discontinuities is rather difficult. In the case of the hollow waveguide with ideal electrically conducting walls, numerous discontinuity problems had been solved using the well known method of orthogonal series expansion. The microstrip waveguide model is suited to apply this method onto microstrip discontinuities, and equivalent circuits with lumped elements can be derived from the measured S-parameters.

The capacitive and inductive element values of the equivalent circuit of most discontinuities are generally very low (often < 0.1pF and < 0.1nH). Within the frequency range of some GHz the parasitic effects may be neglectible. However at higher microwave and millimeterwave frequencies the parasitics strongly influence the transfer characteristics and must be taken into account for the design.

Microstrip discontinuities in the developed circuits can be summarised as:

1. Coaxial cable to microstrip transmissionline transition.
2. Crounding.
3. Components modelleing.
4. Layout modelling.
    a. 90° angle
    b. T-junction
    c. Cross junction

3.3.1 The transition from coaxial-cable to microstrip

The transition from coaxial-cable to microstrip is one of the discontinuities of microwave and millimeter-wave systems, which has normally undesired effects. That is because of the parasitics and also due to the sudden transition from the pure *TEM* mode "The fundamental mode in the coaxial configuration" to the *quasi TEM* mode "The fundamental mode in the microstrip configuration as shown in Fig. 3.4.

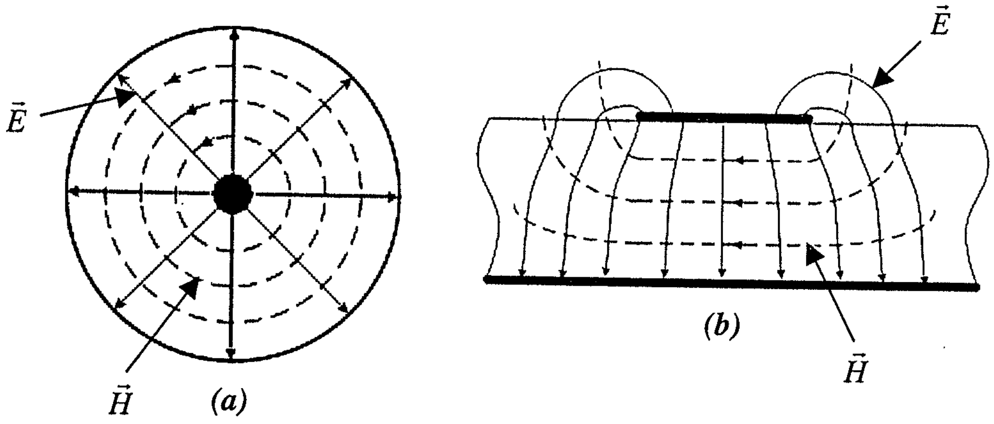

Fig. 3.4 Fundamental modes on coaxial cable and microstrip transmission line. (a) TEM mode in coaxial cable. (b) Quasi TEM mode in microstrip.

Traditionally, the transition is accomplished by aligning the axis of the coaxial connector with the end of the microstrip line, and connecting the outer conductor of the coax to the microstrip's ground plane and the centre conductor to the microstrip as shown in Fig. 3.5.

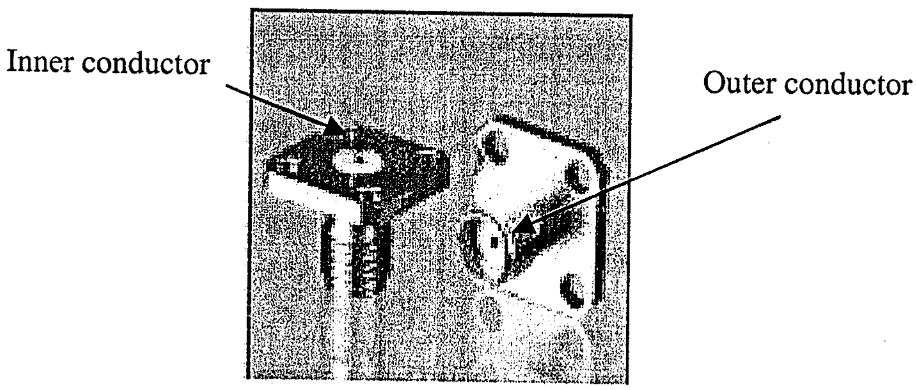

Fig. 3.5 Standard SMA-8 mm connector.

Various improvements have been proposed that extend the transition's bandwidth and enhance its performance, but the basic end-launch geometry has remained the same.