AN ACTIVE SUBHARMONIC RETRODIRECTIVE ARRAY
USING DUAL POLARIZED MICROSTRIP ANTENNAS

by

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ABSTRACT

Retrodirective antenna arrays, when illuminated by an interrogator signal, transmit a signal back towards the interrogator in the same direction as the incoming signal, without any prior knowledge of the source direction. With a retrodirective array the system efficiently transmits a signal without any digital signal processing.

This thesis focuses on the design of an active subharmonic retrodirective array. It involves the design of novel dual polarized coplanar waveguide fed microstrip antenna. Aperture coupling is used for the microwave power to couple to the radiating patch. A common ground plane between the feeds and patch supports the apertures. Orthogonal coplanar waveguide feeds have been used to obtain dual polarization. Also, a part of the thesis deals with the single balanced active subharmonic mixer design. To have a conversion gain, useful in link gain, FETs have been used as the mixing devices. Subharmonic mixer structures are employed when fundamental LO injection is not feasible due to oscillator frequency limitations. A subharmonic mixer is pumped at half the LO frequency, and the RF signal is mixed with the second harmonic of the LO frequency.

Full-wave analyses were performed in the commercial software package Ansoft Ensemble. Subharmonic mixer was realized in Serenade SV. The antenna and mixer models were implemented in hardware as well using a T-Tech milling machine. Results were measured on a network analyzer to show simulation validity.

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CHAPTER 1

INTRODUCTION

Background

In a world of increasing mobility, there is a growing need for people to communicate with each other and have timely access to information regardless of the location of the individuals or the information. The demand for wireless communication systems of increasing sophistication and ubiquity has led to the need for a better understanding of fundamental issues in communication theory and electromagnetics and their implications for the design of highly-capable wireless communication systems.

Antennas are an integral part of a wireless communication system. A well-designed antenna is required for the information signals to be efficiently received and transmitted [1]. The antenna when integrated with active devices such as transistors, mixers, and amplifiers offers improved performance; and an array of antennas coupled with these active devices can be utilized for specific applications in the communication systems. One such antenna array which utilizes microwave mixers to improve its performance is the retrodirective array.

Motivation

Retrodirective antenna arrays have the unique characteristic that when the array is illuminated by a signal, it will transmit the signal towards the source without any prior information on its location [2]. A retrodirective array can be used effectively in a mobile communication system, such as from a ground station to moving vehicles, aircraft, or satellites. Such arrays will play important

roles in future digital mobile communication systems where high link gain and self-beam tracking are desired.

The classic method of achieving retrodirectivity is the Van Atta array, which is illustrated in Fig. 1.1. In this arrangement a sample of the incoming wavefront received by the antenna array elements to the right of the center of the array is retransmitted from the mirror image element on the left side of the array. Each wavefront sample experiences the same time delay, achieved by interconnecting the elements of each mirror image pair by a transmission line, the lengths of the transmission lines being equal.

Fig. 1.1: One-dimensional Van-Atta array

The other method, the more popular of the two, to achieve retrodirectivity uses heterodyne mixing. The basic principle underlying this method is, when the LO frequency is twice the RF frequency, a mixer provides the IF that has the same frequency as the RF and the phase is the conjugation of the RF signal. This phase conjugation is what is needed to reconstruct the outgoing wave. It is essential in this method to suppress RF leakage as the RF and IF frequencies are the same. Researchers have delved into different kinds of passive and active retrodirective arrays [6-12]. The criteria for designing a good retrodirective array

have been good RF/IF and LO/IF isolation, conversion gain from the mixer, LO frequency requirements which is not so high as to increase the system design. This thesis explores the heterodyne mixing technique and comes up with a novel architecture for a retrodirective array. Keeping in view the criteria for better array performance, an optimized array design is presented. A novel dual-polarized antenna is designed with coplanar waveguide feed lines to provide better isolation between the transmitting and receiving ends of the array. The array design also employs the subharmonic mixing using MESFETs which provides the advantage of lower LO frequency requirements coupled with higher conversion gain.

Organization of the Thesis

Chapter 2 is a review of antenna and mixer design. Chapter 3 introduces the retrodirective arrays, their principles, mathematical description and different kinds of arrays. In Chapter 4 a novel design for a dual-polarized microstrip antenna is discussed. The chapter also presents simulation and experimental results. Chapter 5 presents the architecture for a balanced active subharmonic mixer. Chapter 6 provides the description and results of the complete retrodirective array. Chapter 7 summarizes the thesis with conclusions and provides scope for future work.

CHAPTER 2

LITERATURE REVIEW

This thesis can broadly be split into two parts; antenna design and subharmonic mixer design. This chapter gives an introduction to antennas and mixers. The objective is to provide the basics of antennas and mixers which help in understanding the further chapters in the thesis document.

Antenna Parameters

An antenna is an interface between open-space and a guiding device or transmission line; it is a means for radiating or receiving radio waves [1]. An antenna in addition to this function also must serve as a directional device, i.e. optimize or accentuate the radiation energy in some directions and suppress it in others. Some parameters which quantify the performance of an antenna are briefly described below [1]:

1. Radiation Pattern: Mathematical function or graphical representation of the radiation properties of the antenna as a function of space coordinates.
2. Principal Patterns: The E- and H-plane patterns. The E-plane is defined as the plane containing the electric field vector and direction of maximum radiation. The H-plane is the plane containing the magnetic-field vector and the direction of maximum radiation.
3. Radiation Pattern Lobes: A radiation lobe is a portion of the radiation pattern bounded by regions of weak or zero radiation intensity. The main lobe contains the direction of maximum radiation. A radiation lobe whose axis makes an angle approximately 180° with respect to the main lobe of the antenna is called a back lobe. Other lobes in the pattern are termed side lobes.
4. Radiation Intensity: The power radiated from an antenna per unit solid angle.

5. Directivity: The ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions.
6. Gain: The ratio of the radiation intensity, in a given direction, to the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically (radiated equally in all directions).
7. Polarization: it describes the time varying direction and relative magnitude of the electric-field vector; specifically, the figure traced as a function of time by the extremity of the vector at a fixed location in space.

Dual-Polarized Antennas

Wireless communications systems are prone to interference and poor signal quality due to omnidirectional transmission and reception of signals. One technique to improve this degradation is polarization diversity. The use of polarization diversity, particularly dual-polarized antennas, is a promising cost-effective and space-efficient alternative where two spatially separated uni-polarized antennas are replaced by a single antenna structure employing orthogonally polarized elements.

Or the dual polarization can be used, as in this thesis, where there is a need to double the number of channels available in a limited bandwidth. In the thesis, one polarization of a microstrip antenna is utilized for receiving the signal and the other for transmitting it. Many researchers working with microstrip antennas have delved into the aperture coupling feed technique, with its inherent advantages to realize dual polarized antennas, and this technique is used in this thesis [3].

Some useful features related to aperture coupled microstrip antennas are:

- Independent selection of antenna and feed substrate materials
- Two-layer construction shields radiating patch from feed network

- Increased substrate space for antenna elements and feed lines
- Convenient integration for active arrays
- Low cross polarization in principle planes
- Variations in patch shape, aperture shape, feed line type, etc.
- Extension to aperture-coupled microstrip line couplers, waveguide transitions, dielectric resonators, etc.

A prototype of an aperture-coupled antenna is shown in Figure 2.1 [3].

Fig. 2.1: An aperture coupled antenna prototype [3]

Different shapes of the radiating element have been investigated in the literature [3], like the square patch, circular patch, star patch etc. Square patches may result in the generation of high cross polarization levels, and thus should be avoided unless dual or circular polarization is required. Generally, for dual polarization the designs consist of the use of either two centered crossed slots or

two offset orthogonal slots cut on the ground plane to excite two orthogonal modes. Many researchers have presented different configurations of dual polarized aperture coupled antennas since their appearance in the 1980s. Fig. 2.2 shows one such example [3]. Here, port 1 and port 2 are the two orthogonally fed ports, L_{a1} , L_{a2} are the lengths of the slots (coupling apertures) in the ground plane, W_a the width of the slot (coupling aperture) in the ground plane, and R_p is the radius of the circular patch (radiating element).

Fig. 2.2: Dual-polarized microstrip fed antenna [3]

Fig. 2.3 shows the radiation pattern in the case of port 1 excitation. The figure shows the cross polarization level of less than -20 dB has been obtained. The same result was obtained for port 2 excitation.

Fig. 2.3: Radiation pattern of the dual polarized antenna in Fig. 2.2 [3]

A new dual-polarized antenna to compensate for the shortcomings of the antennas presented in the past by researchers [6, 7, 13]. In the past, there had been a compromise between the bandwidth the antennas provided and the number of patches used. The type of feed line and the isolation between the two-ports were also a matter of concern. Hence, adapting to the advantages provided by aperture coupling and conductor backed coplanar waveguide type feed (will be discussed in Chapter 5); an antenna was built to provide better results.

A major part of this thesis deals with construction of a balanced subharmonic mixer; hence it becomes essential to discuss mixer basics and the terminology used in mixer design.

Mixers

Mixers are frequency-translation devices and are critical components in modern radio-frequency (RF) systems. A mixer converts RF power at one frequency into power at another frequency to make signal processing, like

amplification of the received signal at a frequency other than the RF, improved selectivity (filtering) and an easier implementation of low noise, easier and less expensive [4].

Fig. 2.4: Block diagram of an ideal mixer

The ideal mixer, represented by Fig. 2.4, is a device which multiplies two input signals. If the inputs are sinusoids, the ideal mixer output is the sum and difference frequencies given by

$$V_o = [A_1 \cos(\omega_1 t)][A_2 \cos(\omega_2 t)] = \frac{A_1 A_2}{2} [\cos(\omega_1 - \omega_2)t + \cos(\omega_1 + \omega_2)t] \quad \text{Eq. (2.1)}$$

Typically, either the sum, or the difference, frequency is removed with a filter. In reality, mixers produce more than just the sum and difference frequencies. The formula for all the frequencies can be written as $IF = n \cdot RF \pm m \cdot LO$ (n, m are integers). The non-linear behavior of a semiconductor device is used to realize the mixing function, which can be described by Eq. 2.2. Diodes, field effect transistors (FET) and bipolar transistors can all be used in mixer designs. Fig. 2.5 shows the typical I-V characteristics of a Schottky diode. Most modern diode mixer designs use Schottky diodes as they are majority carrier

device which means they have a higher switching speed than p-n junction diodes [4].

$$I = a_0 + a_1V + a_2V^2 + a_3V^3 + \dots \quad \text{Eq. (2.2)}$$

Fig. 2.5: I-V characteristics of a Schottky diode

Diodes are nonlinear devices, with a strong a_2 component (see Eq. 2.2). The other components have low amplitudes or can be distilled using a filter. Thus, the basic mixer design entails injecting the signals to be mixed and extracting the desired mixing product whilst maximizing the efficiency of the conversion. The diode is biased to have a quiescent operating point close to the region of maximum non-linearity in its operating characteristics which allows the diodes square law characteristic to be traversed with lower levels of LO drive.

Suppose the RF and LO signals are represented as follows:

$$V_{RF}(t) = V_{RF}\cos(2\pi\nu_{RF}t + \phi_{RF}) \quad \text{Eq. (2.3)}$$

$$V_{LO}(t) = V_{LO}\cos(2\pi\nu_{LO}t) \quad \text{Eq. (2.4)}$$

The mixer superimposes the local oscillator signal with the RF signal, a non-linear element (such as a diode) whose output is

$$I(t) = a_0 + a_1(V_{RF}(t) + V_{LO}(t)) + a_2(V_{RF}(t) + V_{LO}(t))^2 + a_3(V_{RF}(t) + V_{LO}(t))^3 + \dots \quad \text{Eq. (2.5)}$$

The second-order (quadratic) term of Eq. 2.3 can be expressed as

$$I(t) = a_2 V_{RF}^2 \cos^2(2\pi\nu_{RF}t + \phi_{RF}) + a_2 V_{LO}^2 \cos^2(2\pi\nu_{LO}t) + 2a_2 V_{RF} V_{LO} \cos(2\pi\nu_{LO}t) \cos(2\pi\nu_{RF}t + \phi_{RF}) \quad \text{Eq. (2.6)}$$

Developing the product of the two cosine functions, we obtain

$$I(t) = \dots a_2 V_{RF} V_{LO} \cos(2\pi(\nu_{LO} + \nu_{RF})t + \phi_{RF}) + a_2 V_{RF} V_{LO} \cos(2\pi(\nu_{LO} - \nu_{RF})t - \phi_{RF}) \dots \quad \text{Eq. (2.7)}$$

There are obviously other terms in $2\nu_{LO}$, 2ν , $3\nu_{LO} + \nu$, $3\nu_{LO} - \nu$ etc. in the above equation, as well as terms at very different frequencies like ν , 3ν etc. By inserting a filter at the output of the mixer, the unwanted products can be discarded.

Hence, after mixing and filtering, the output is

$$I(t) \propto V_{RF} V_{LO} \cos(2\pi(\nu_{LO} - \nu_{RF})t - \phi_{RF}) \quad \text{(Eq. 2.8)}$$

Eq. (2.8) shows the signal will have a change in frequency from ν_{RF} to $(\nu_{RF} - \nu_{LO})$ or $(\nu_{LO} - \nu_{RF})$, will be proportional to the original electric field of the incident wave (V_{RF}) and a phase relation $(-\phi_{RF})$.

A mixer's performance is expressed by the following parameters:

Conversion gain is the ratio of the output signal level to the input signal level expressed in dB.

Isolation is the amount of "leakage" or "feedthrough" between the mixer ports.

Noise Figure is the signal-to-noise ratio at the input divided by the signal-to-noise ratio at the output expressed in dB.

Dynamic Range is the amplitude range over which a mixer can operate without degradation of performance. The 1 dB compression point is generally taken to be the top of the dynamic range of a mixer because the input RF power that is not converted into desired IF output power is instead converted into heat and higher order intermodulation products. The intermodulation products that begin to appear when RF power is increased beyond the 1 dB compression point can begin to obscure the desired IF output. The lower bound of the dynamic range is taken to be the noise level of the mixer.

Figure 2.6 is a typical single balanced mixer block diagram. The mixer diodes are reversed relative to each other; the desired frequency (IF) components of each diode are then in-phase while the DC outputs are positive and negative respectively. The two diode outputs are summed in a tee where the DC terms cancel and only the desired IF component exists at the IF port [4].

Fig. 2.6: Single balanced mixer block diagram

FET Mixers

FET mixers are, mostly, transconductance mixers using the LO signal to vary the transconductance of the transistor. They have the advantage of providing the possibility of conversion gain rather than loss and can also have lower noise figures than passive designs. Fig. 2.7 shows a simple transconductance FET mixer. The RF (and LO) short circuit at the drain is important to ensure that the value of V_{ds} (drain to source voltage) is not moved significantly from its DC bias point by the applied LO. This ensures the magnitude of the time varying transconductance is maximized so optimizing the conversion gain.

Fig. 2.7: A transconductance mixer

This thesis involves the use of FETs in subharmonic mixer design. They provide good conversion gain, which can be useful to improve the link range in a communication system. The details will be discussed in Chapter 5.

CHAPTER 3

REVIEW OF RETRODIRECTIVE ARRAYS

A retrodirective antenna array transmits a signal back to the interrogator's position without any prior knowledge of the incoming angle or relying on sophisticated digital signal processing algorithms [2]. This kind of an array performs beam-steering action automatically by virtue of the array architecture used. The basic principle on which a retrodirective antenna array operates relies on the self conjugation of an incoming wavefront, i.e. each element in the array must re-radiate a received signal, with the phase of the transmitted signal acting as the conjugate of the received signal. Each element in the array is independently phased; this phasing is derived directly from the signal each array element receives. This self-phasing behavior compensates for the inhomogeneities in the propagation characteristics of the environment through which the signal is traveling. These kinds of arrays find applications in advanced digital mobile communication systems where high link gain and self-beam tracking are desired. They can be effective in communication with moving vehicles like aircraft and satellites.

Types of Retrodirective Arrays

Retrodirective arrays can broadly be classified into the following kinds:

- (i) Corner reflector,
- (ii) Van-Atta array,
- (iii) Heterodyne retrodirective arrays.

Corner Reflector

This is the simplest retrodirective structure, which shows the principle of retrodirectivity or in general phase reversal. Fig. 3.1 shows a corner retrodirective reflector [6]. This reflector consists of two flat conducting plates set at an angle of 90° to each other. An incident signal arriving at an angle θ with respect to plate 'Y' will be reflected back at its Snell angles as shown in the Fig. 3.1. The reflected signal from plate 'Y' will arrive at an angle $90^\circ - \theta$ with respect to the plate 'X' and will again be reflected at its Snell angle. Here the angle of this reflected signal with respect to $a-a'$, which is parallel to the plate 'Y', is the same as the angle of the incident wave with respect to the plate 'Y'. This indicates that the outgoing signal is returned in parallel to the incoming signal, i.e. returns back in the direction from which it arrived.

Fig. 3.1: Corner reflector

Van-Atta Array

In this array, samples of the incoming wavefront received by the antenna array elements to the left of the center of the array are retransmitted from the mirror image element on the right side of the array. Each wavefront experiences the same delay, achieved by interconnecting the elements of each mirror image pair by a transmission line, the lengths of the lines being equal. Fig. 3.2 shows a simple Van Atta array.

Fig. 3.2: A 1-D Van Atta array

In Fig. 3.3, if AF represents an incoming wavefront, given that the total path lengths are the same, i.e.,

$$AA' + FF' = BB' + EE' = CC' + DD',$$

then by introducing a delay τ , same for all the paths, into the connecting paths the retransmitted wavefront will be set up along the same line AF as the incoming wavefront. In general, whatever the direction of the incoming wave, the array is automatically phased correctly for the retransmission of the incoming signal in the same direction as that from which it arrived.

Fig. 3.3: Operation of Van Atta array

If scattered energy is neglected then the resultant retransmitted field from the Van Atta array, E_{rt} , can be expressed as

$$E_{rt} = A \times e^{j(\omega t + \phi_L)} \sum_{n=-\frac{N}{2}}^{\frac{N}{2}} e^{\frac{j 2 \pi x_n}{\lambda_0} (\sin \theta_t - \sin \theta_r)} \quad \text{Eq. (3.1)}$$

Here,

ω = array operating frequency in radians per second, Φ_L = interconnecting line phase delay in radians, θ_r = angle of arrival in radians measured from array broadside, θ_t = angle of retransmission in radians measured from array broadside, and x_n = spacing in meters of the n th receive element from the array phase centre. If for the above conditions $\theta_r = \theta_t$, retrodirective action occurs and the retransmitted wavefront arrives in phase at its point of origin. The choice of ϕ_L is completely arbitrary.

Fig. 3.4 shows an active Van Atta array. In this arrangement circulators are used to separate the signals into forward and reverse transmission paths. These signals are amplified for the retransmission so that the link range can be increased. The major problem with this kind of arrangement is the introduction of differential phase errors in the retransmitted signal due to the amplifiers and circulators.

Fig. 3.4: Array pair in an active Van Atta array

The Van Atta array has a disadvantage, the phase errors lessen the array gain and increase the array beam width and side lobe level [15].

Heterodyne Retrodirective Arrays

In 1969, Pon introduced the principle of heterodyne retrodirectivity where he used phase conjugation with a signal from a local oscillator at twice the incoming signal's frequency [2]. The received signal at each element is phase conjugated by the phase conjugating circuitry. These signals then form the main beam towards the source direction. The most essential part of the heterodyne

phase conjugation technique is a frequency-mixing device (mixer). Devices with non-linear behavior can be used as mixing devices, as discussed in Chapter 2. Passive device mixers suffer conversion loss but can be operated at low LO power. But active device mixers which yield conversion gain with very little LO power form the best solution for a simple low cost retrodirective array.

Phase conjugation with heterodyne mixing uses an LO signal at twice the RF frequency. In this scheme, the lower sideband (difference frequency) product has the same frequency as the RF, but with a conjugated phase. Example of one such array is shown in Fig. 3.5. In practice, a small frequency offset between incident and retransmitted waveforms is introduced to facilitate frequency discriminations of incoming and outgoing signals at the detection apparatus.

Fig. 3.5: Array pair in a heterodyne mixing retrodirective array [12]

In order to see how the conjugate signal can be generated at each element using microwave circuitry, consider Equations 2.3, 2.4 and 2.8 in Chapter 2:

$$V_{RF}(t) = V_{RF} \cos(2\pi\nu_{RF}t + \phi_{RF})$$

$$V_{LO}(t) = V_{LO} \cos(2\pi\nu_{LO}t)$$

$$I(t) \propto V_{RF}V_{LO} \cos(2\pi(\nu_{LO} - \nu_{RF})t - \phi_{RF})$$

Now consider a $2\nu_{RF}$ signal (two times the RF frequency) delivered to the LO port of the mixer. This $2\nu_{RF}$ LO signal has to be delivered to all elements with the same amplitude and phase; otherwise the mixed output will contain spurious outputs. If this ever happens, the sum of the excited field at each element will be distorted and will not form the conjugate beam. Using difference frequency generation in the mixer, the IF output current can be written as:

$$I(t) \propto V_{RF}V_{LO} \cos(2\pi(\nu_{RF})t - \phi_{RF}) \quad \text{Eq. (3.2)}$$

This current component has the conjugate phase $-\phi_{RF}$ instead of the input phase $+\phi_{RF}$. Therefore when it is delivered to the antenna, it will excite the field which forms the phase-conjugate wave. This wave will propagate backwards with the same wavefront as those of the incoming wave. By using amplifiers after the mixing process at each element, the conjugate beam will be substantially more powerful than the initial input signal. Another interesting conclusion is that a phase-conjugate antenna array can have any shape and can be placed on any surface as long as the $2\nu_{RF}$ LO signal is delivered to all elements in phase. This feature can be very useful for vehicular communication systems because the antenna arrays can be readily designed to accommodate the shapes of the vehicles.

There has been a lot of work done to improve the basic heterodyne retrodirective arrays, to fit the application and requirement [6-12]. For example, many researchers have worked on the passive retrodirective arrays, and some in active counterparts. Variations in the type of mixers and the antennas have

played an important role in the development [6-12]. The following section explains several kinds of retrodirective arrays that have appeared in the literature.

Retrodirective Antenna Array Using Power Dividers

One of the basic forms of retrodirectivity was demonstrated by Karode and Fusco using power dividers [7] (see Fig. 3.6). Phase conjugation is achieved with a LO frequency of 1.99 GHz (twice the incident). The mixer output at 1 GHz is amplified and sent back to the microstrip antenna with the help of a power divider. The power divider acts as a circulator to provide isolation between the incident and the retransmitted signal. But the power divider comes with a 3 dB loss. The antenna here provides a low bandwidth. The radiation pattern in Fig. 3.7 shows retrodirectivity.

Fig. 3.6: Retrodirective array using power divider [7]

Fig. 3.7: Antenna pattern for a retrodirective antenna using a power divider [7]

Retrodirective Antenna Array Using Dual-Polarized Patch

The circuit for this configuration is similar to Fig. 3.6, except that the polarization properties of the antenna are used. The microstrip antenna and the power divider are replaced with a dual polarized antenna. The two linear orthogonal modes TM_{10} and TM_{01} are used to provide dual polarization. One mode is used for receiving and the other for transmitting. Isolation between the modes ensures the isolation between the received and retransmitted waves. The modes are operated at approximately the same frequency. The advantage of this arrangement is lower path loss, but it also has poorer isolation. The kind of antenna which we use should be able to work in orthogonal modes. Fig. 3.8, Fig. 3.9, and Fig. 3.10 below describe the circuit and the results.

Fig. 3.8: Retrodirective antenna using dual-polarized patch antenna [7]

Fig. 3.9: Radiation patterns of dual-polarized patch antenna [7]

Fig. 3.10: Antenna pattern for retrodirective antenna using dual-polarized patch antenna [7]

Balanced Quasi-Optical FET Mixers With Conversion Gain

A retrodirective array with an intrinsic gain was designed by Miyamoto et al. [9]. A balanced FET mixer was used which provides the gain and also an isolation of 20 dB between RF-IF. As the RF and IF frequencies are very close it is impossible to use conventional filters and a hybrid (dual-frequency rat race coupler, shown in Fig. 3.11) with phase delay was employed. This provides an effective RF/IF isolation.

Fig. 3.11: Dual frequency hybrid (rat race coupler)

Fig. 3.12 shows the retrodirective array (1 element). A dual-frequency rat race with a common output port is realized. The RF and LO leakage are out of phase and the IF signal is in phase at the output ports. As a result, the RF and LO leakage can be cancelled out and a combined IF is obtained. The phase conjugating element consists of a balanced quasi-optical mixer with an integrated dual polarized antenna. The FET mixers provide conversion gain. Over 30 dB isolation between the RF and LO ports was measured. Since both the RF and LO signals applied to each FET are out of phase, the leakage is cancelled when

combined at the output port, and the IF signals are in phase and hence add up. Also, RF-IF isolation of 20 dB was achieved at the output.

Fig. 3.12: Phase-conjugating element using a balanced mixer with an integrated patch antenna [9]

Balanced Harmonic Mixer

A balanced version of a subharmonic mixer was built by Brabetz and Fusco [11]. Subharmonic mixer structures are employed when fundamental LO injection is not feasible due to oscillator frequency limitations, i.e., when sources with high enough frequency are not readily available due to economics or other system design constraints. Any mixer structure can be operated at a subharmonic of the LO, but the conversion efficiencies tend to be poor unless the mixer is specifically designed to extract a particular mixing product. The structure has a rat race coupler to whose DIFFERENCE port the LO signal is applied, whereas the RF signal is applied to the SUM port. Two pairs of anti-parallel Schottky diodes are used as mixers and are connected to the other two arms of the hybrid. Fig. 3.13 shows the layout used to build the subharmonic mixers. The LO/IF isolation at the output port is provided by the structure of the hybrid; the LO signal is cancelled since the LO signals from both the mixers are 180° out-of-

phase. The LO drive frequency in this arrangement is eased, which, until now, had to be at twice the RF frequency.

Fig. 3.13: Layout of the balanced subharmonic mixer [11]

The phase conjugation action is achieved with RF and LO frequencies approximately equal. LO/IF isolation of more than 29 dB was found at the required frequency band as shown in Fig. 3.14.

Fig. 3.14: LO/IF isolation for layout in Fig. 3.13 [11]

The measured response of a two-element retrodirective array is shown in Fig. 3.15. The plot of the E-field maximum of the transmitted signal is measured, which has flatter azimuthal response than its passive counterpart.

Fig. 3.15: Two-element Retrodirective-array response [11]

The above example of a retrodirective antenna using subharmonic mixing employs passive devices (Schottky diodes) which result in conversion loss. In this thesis, we extend the above principle of subharmonic mixing to active devices (FETs), resulting in conversion gain instead of loss. Furthermore, a new conductor backed coplanar waveguide fed dual polarized antenna has been employed.

CHAPTER 4

DUAL-POLARIZED PATCH ANTENNA

This chapter introduces a novel architecture for a dual-polarized microstrip antenna. The design stems from the idea of previous aperture-coupled antennas [12–15] and involves a conductor backed coplanar waveguide fed patch antenna. The chapter also mentions some advantages of choosing the topology and the considerations made for the design of the antenna.

Patch antennas are inviting wide interest in various mobile communication systems since they can provide advantages over traditional whip and helix antennas in terms of high efficiency, polarization diversity for enhanced reception and transmission functions, and increased mechanical reliability. They also find applications in Bluetooth which provides short-range wireless communication between electronic devices such as PCs and cell phones.

Researchers have been utilizing many kinds of feed lines to excite the patch antennas, the most prominent being microstrip transmission line and coplanar waveguide. Coplanar waveguide (CPW) is utilized in many commercial microwave circuits where uniplanar fabrication is required. A conventional CPW on a dielectric substrate consists of a central strip conductor with semi-infinite ground planes on either side. This structure supports a quasi-TEM mode of propagation. A cross section of a conventional CPW and the quasi-TEM E-field is shown in Fig. 4.1 [13].

Fig. 4.1: Conventional coplanar waveguide and quasi-TEM mode

With the signal and ground plane on the same surface, it is convenient to mount active and passive devices on the same side of the substrate and eliminate the need for vias commonly used with microstrip transmission lines. CPWs also provide low insertion loss and lower dispersion when compared to their microstrip counterparts.

Due to mechanical and thermal requirements, some substrates for coplanar monolithic microwave integrated circuits (CMMIC) are also metalized on one side. The resulting transmission line is referred to as conductor-backed CPW (CBCPW) [13]. The metalized layer (ground plane) provides mechanical strength for a thin and fragile substrate and acts as a heat sink for circuits with active devices. A CBCPW with ground planes assumed infinite in the lateral directions is shown in Fig. 4.2.

Fig. 4.2: Conductor backed coplanar waveguide

Dual-polarized microstrip antennas have been proposed by Adrian and Schaubert [3]. It was shown that orthogonal feeds and cross slots architecture can

yield dual polarization. The idea was illustrated in Fig. 2.2. Examples of an aperture-coupled microstrip antenna fed by coplanar waveguides have been proposed by Lee and Simons [14]. A dog-bone shaped aperture is etched in the common ground plane. The feed network and the radiating patch are fabricated on different substrates. The microwave power is coupled from the feed lines to the patch through the aperture. The above proposed designs have shortcomings in the form of low bandwidth or poor isolation for dual-polarized antennas.

In order to improve the bandwidth and isolation, we investigated a CBCPW fed patch antenna. To compare the performance of the proposed antenna with the conventional CPW fed antenna, we simulated, in *Ansoft Ensemble*, a microstrip antenna with two orthogonal coplanar waveguide feed lines. The layout of the antenna is shown in Fig. 4.3. A 50 mil RT/Duroid with $\epsilon_r = 10.2$ was used as the dielectric substrate. A full-wave analysis was performed on the design, the scattering matrix results of which are shown in Fig. 4.4. The $|S_{11}|$ (return loss) was found to be -9 dB and $|S_{22}|$ (return loss) was found to be -4 dB. The $|S_{21}|$ (isolation) between the ports is -5 dB. This shows that, a perfect match was not obtained in the desired frequency range and also the isolation was poor.

Fig. 4.3: View of patch antenna with orthogonal coplanar waveguide feed lines

Fig. 4.4: Scattering matrix with orthogonal CPW feeds

The proposed antenna design in this thesis consists of two separate dielectrics, one for the feed network and the other for the radiating patch. The two substrates are separated by a common ground plane which contains the

apertures that electromagnetically couple the microwave power to the patch as shown in Fig. 4.5. The full wave analysis of the design was performed using *Ansoft Ensemble*. The antenna was optimized to perform in the 3.0-3.5 GHz band.

Fig. 4.5: An exploded view of the dual-polarized CBCPW antenna

The choice of dielectric substrates took the following points under consideration:

- The antenna substrate dielectric constant affects the bandwidth and radiation efficiency of the antenna, with lower permittivity giving wider impedance bandwidth and reduced surface wave excitation.
- The antenna substrate thickness affects bandwidth and coupling level. A thicker substrate results in wider bandwidth.

- c. Thinner feed substrate result in less spurious radiation from feed lines, but higher loss. A compromise of 0.01λ to 0.02λ is usually good.

With these considerations in mind and the availability of the substrates, the lower or feed substrate was chosen to be a 62 mil RT/Duroid with $\epsilon_r = 2.33$ and the upper or antenna substrate a 50 mil RT/Duroid with $\epsilon_r = 10.2$.

Fig. 4.6: Dimensions for the CBCPW for a characteristic impedance of 50 ohms

The feed network in the lower substrate consists of two orthogonal CPW feeds. Using the *Ensemble Estimator*, the dimensions for the CPW feed line were found. The feed lines were designed for characteristic impedance $Z_0 = 50$ ohms. To come to an agreement with the values found using the *Estimator* a few iterations were run in *Ensemble*. Fig. 4.6 shows the CPW waveguide with its dimensions for our frequency band. The simulated port impedance is shown in Fig. 4.7.

Fig. 4.7: Simulated characteristic port impedance for a CBCPW

An exploded view of the antenna with its layers and the common ground plane is shown in Fig. 4.5. The coupling takes place from the slot at the end of the CPW feed. The feed terminations can either be a modified short circuit or an open circuit. The short circuit gives rise to excessive inductance at the termination and the coupling from an open circuit is capacitive. We have employed the short circuit termination for the feed line, which has been proved by researchers to have a better bandwidth than its open circuit counterpart [13]. It also eases the design considerations. A view of the feed network with the dimensions is shown in Fig.4.8. In order for the arms of the feeds not to intersect each other, Port 2 has been offset by a distance of 7 mm from the center of the patch. The lengths of the slots were optimized for matching and isolation in the 3.0-3.5 GHz band.

Fig. 4.8: Feed network on the lower substrate

Two apertures are made in the common ground plane to allow the coupling of energy to the patch. They are placed exactly on top of the slots of the feed lines on the other side of the substrate. These apertures result in back radiations, though not significant, which is an inherent property of aperture coupling. The apertures are made wider than the feed line slots to allow easier alignment during the assembly of the fabricated antenna. The apertures are exactly the same length as the slots. Fig. 4.9 shows the apertures in the ground plane.

Fig. 4.9: Apertures (slots) in the common ground plane

A square patch on the top substrate is used as the radiating element in the antenna. The patch size was optimized with iterations in *Ensemble* to $24 \times 24 \text{ mm}^2$ (shown in Fig. 4.10).

Fig. 4.10: Radiating patch on the top substrate

The full-wave analysis was performed in *Ensemble* of the structure shown in Fig. 4.5. The scattering matrix is shown in Fig. 4.11.

Fig. 4.11: Scattering matrix of the dual-polarized antenna (simulated in *Ansoft Ensemble*)

When compared to the results in Fig. 4.4, better match and isolation are obtained in the desired band of 3.0–3.5 GHz. The simulated results show the antenna resonates at a frequency of 3.18 GHz and provides an impedance bandwidth of 15% (10 dB criterion), which is quite large for a microstrip antenna. The return loss ($|S_{11}|$) for the port 1 was found to be -18.5 dB and for port 2 more than -35 dB, which are much lower than the -10 dB criterion. Also a good isolation ($|S_{21}|$) of -25 dB was found between the two ports. Fig. 4.12 shows the VSWR of the designed antenna.

Fig. 4.12: Dual-polarized antenna VSWR

The main purpose of the antenna design was to provide dual-polarization. When only one port is excited, the radiation pattern produced is essentially linearly polarized parallel to the feed line for that port. The radiation patterns at 3.18 GHz are shown in Fig. 4.13 and Fig. 4.14. The cross-polarization level is found to be 20 dB less than the co-polarized level. The figure compares the electric field patterns (both theta and phi components) at $\varphi = 0^\circ$ and $\varphi = 90^\circ$. The front to back radiation is better than 10 dB.

(a)

(b)

Fig. 4.13: Far-field radiation pattern for port 1 excitation at 3.18 GHz
 (a) $\phi = 0^\circ$ (b) $\phi = 90^\circ$

(a)

(b)

Fig. 4.14: Far-field radiation pattern for port 2 excitation at 3.18 GHz
(a) $\phi = 0^\circ$ (b) $\phi = 90^\circ$

Limitations of the manufacturing process like hand soldering of metal vias (copper wire) and SMA connectors can cause discrepancies between the return loss of the simulated and fabricated antenna. Also, an air gap between the two dielectrics, due to small amounts of solder, can cause deviation from the expected results. Hence, an additional air dielectric layer (thickness 32 mil, $\epsilon_r = 1$) was incorporated in the *Ensemble* design of the antenna and another full-wave analysis performed. The feed lines were optimized for 50 ohm characteristic impedance. Fig. 4.15 shows the scattering matrix results for the new design.

Fig. 4.15: Scattering matrix of dual-polarized antenna with additional air dielectric

Fig. 4.15 shows that if an additional air dielectric is added, a shift in the resonance occurs. The resonance frequency here is 3.8 GHz. Hence, it can be stated that the extra solder can cause undesired results.

To verify the validity of the simulation results the antenna was fabricated using a T-Tech milling machine. The substrates used were 62 mil with $\epsilon_r = 2.33$ for the radiating element and a 50 mil with $\epsilon_r = 10.2$ for the feed lines.

Further, care must be taken to ensure that the CPW supports only a quasi-static TEM mode. The geometry of a conductor-backed CPW may cause strange effects since an additional fundamental mode, the parallel-plate line mode (PPL), see Fig. 4.16, appears. This PPL mode is closely related to the voltage difference between the ground conductors on the top of the substrate and on the bottom. A particular disadvantage of the PPL mode is that undesired coupling between different parts of a circuit becomes possible. Hence, the topside of the coplanar ground planes are connected to the bottom ground planes by filled metal vias. The vias short circuit the electric field of the parasitic PPL mode and thus suppress its propagation.

Fig. 4.16: Parallel-plate mode (PPL)

As the desired results were not obtained in the laboratory, the design would be sent to a company for building. With the modern techniques of building and lesser probability of error, we hope the antenna design should work as expected from the simulated results.

CHAPTER 5

DESIGN OF BALANCED SUBHARMONIC MIXER

Background

As discussed in chapter 2 (section 2.3); Mixers are frequency translation devices. They allow the conversion of signals between a high frequency (RF) and a lower Intermediate Frequency (IF) with the use of fixed Local oscillator frequency (LO). In communications systems, the RF is the transmission frequency, which is converted to an IF to allow improved selectivity (filtering) and an easier implementation of low noise and high gain amplification. The parameters characterizing the mixer performance were discussed briefly in Chapter 2.

The non-linear behavior of semiconductor devices, like Diodes, Field Effect Transistors (FETs) and heterojunction bipolar transistors, is used to realize the mixing function, as shown in Chapter 2. Passive devices, like diodes, provide a conversion loss; whereas the active devices like the FETs provide conversion gain, if matched properly at the input and output.

Single FETs or diodes can be used as mixers. However, designs sometimes use combinations of multiple diodes in a balanced structure. Inherent RF - LO isolation can be obtained by using balanced structure. Also, the input power is divided between the multiple devices, the power-handling capacity of a balanced mixer is better than the single device mixer. Single balanced mixers usually use two devices, and are realized as two single-device mixers connected by a 180- or 90-deg hybrid [4]. Image rejection filters also use the hybrids to cancel the image, which has a frequency above or below the local oscillator (LO) frequency by an

amount equal to the IF frequency, which is an unwanted input signal to the mixer.

Subharmonic mixer structures are employed when fundamental LO injection is not feasible due to oscillator frequency limitations. A subharmonic mixer is pumped at half the LO frequency, and the RF signal is mixed with the second harmonic of the LO frequency. In a “standard” mixer, the product of interest is $LO \pm RF$, but for a subharmonic mixer, $2LO \pm RF$ is the desired signal. The fundamental conversion components $LO \pm RF$ are out of phase at the output, so the devices effectively short - circuit each other at the fundamental mixing frequency. Traditionally, subharmonic mixers use an antiparallel Schottky diode pair to generate a nonlinear conductance waveform. But, Maas [4] has shown that FETs also can be used in design of subharmonic mixers.

This chapter details the design of an active balanced subharmonic mixer and also proposes ideas to achieve better RF-IF isolation when used in design of retrodirective arrays.

Mixer Design

The subharmonic mixer involves the use of two FETs as the mixing devices. A subharmonic mixer involves the LO signal being fed out of phase to the gates of the FETs, with the help of a rat-race coupler. The RF signal is fed in phase to the FETs. The IF is the added product from the outputs from the FETs.

Fig. 5.1: Subharmonic mixer schematic

The LO voltages at the gates of the two FETs is out of phase. This is done using a rat race coupler. As seen in Fig. 5.1, the LO signal is fed to the difference port of the rat race coupler and the RF signal fed to the sum port; at the output of the coupler we have two equally split LO, but with opposite phase, and equally split RF.

The following sections describe the subharmonic mixer design and the individual component simulation in *Ansoft Ensemble*.

The FET mixer is a transconductance mixer where the time-varying transconductance is the dominant contributor to frequency conversion and the effect of other nonlinearities are minimal. In designing such a mixer, we need to maximize the fundamental LO-frequency component of the transconductance waveform and minimize the time-variation of other circuit elements like the drain-to-source conductance. The transconductance of a FET is maximum when its drain-to-source voltage V_{ds} is maximum. It is imperative that V_{ds} is not allowed to decrease substantially from its dc value during any part of the LO cycle. Hence, V_{ds} should not drop below the knee of the $I_d(V_{ds})$.

A balanced subharmonic mixer (as seen in Fig. 5.2) is simulated in *Serenade SV*. *Ansoft Ensemble* does not model active devices; hence *Serenade SV* was chosen to perform analysis on the subharmonic mixer.

Two N-Channel GaAs MESFET NE722S01 (courtesy: California Eastern labs) are used as the mixing devices. The data sheets of the MESFET are provided in Appendix I. The FET TOM model (Appendix B) provided in *serenade SV* has been used. The TOM model data specifications provided by the manufacturer were incorporated in the FETs. The transistors are biased at $V_{DS} = 3\text{ V}$ and $V_{GS} = -0.5\text{ V}$, above the knee of the $I_d(V_{ds})$ curve .

Fig. 5.2 shows the Serenade file schematic of the subharmonic mixer. A LO signal of 3.2 GHz and RF of 3.15 GHz was chosen (a small frequency offset to facilitate frequency discriminations of incoming and outgoing signals at the detection apparatus). It may also be noted that two high value capacitors, at the input, are used to block the dc from entering the rat race coupler. Also a high value capacitor is used to block the dc at the output.

The LO signal is cancelled since the LO signals from the mixers are 180-deg out-of-phase, thus providing high LO-IF isolation. The RF leakage signals through the FETs, however, being in-phase, add at the output, consequently, RF-IF isolation is poor.

Fig. 5.2: Balanced Subharmonic Mixer (Serenade SV schematic)

A regular analysis is performed on the circuit shown in Fig. 5.2, with LO power level of 17.5 dBm and RF power level of -40 dBm. The maximum input power rating (at room temperature) for the NE722S01 is given as 40 mW (16

dBm). The input power is distributed among the two FETs due to the presence of the rat race. Hence, the input power at the FETs would be 3 dB less than the 17.5 dBm applied at the LO port. This makes it a safe power level to operate. Fig. 5.3 shows the reports of the regular analysis on the subharmonic mixer.

Fig. 5.3: Subharmonic mixer spectrum

The results clearly indicate the subharmonic mixing operation. The IF is found to be at 3.25 GHz ($2 \times 3.2 \text{ GHz} - 3.15 \text{ GHz}$), $2 \times \text{LO} - \text{RF}$. The IF level is found to be -32 dBm. The conversion gain is defined as ratio of the output IF signal level to the input RF signal level expressed in dB. In this case, we see a gain about 7 dB. Hence, the use of FETs provides us with a gain of 7 dB, which is better than the conversion loss provided by the passive devices like the diodes. The feeding of out of phase LO signal to the FETs achieves very good LO-IF

isolation of 80 dB. The fundamental mixing product level at 50 MHz is very low at -91 dBm. But, the RF-IF isolation is very poor, which can be improved by using other techniques discussed in the later sections of this chapter.

A sweep analysis is performed with the LO power level swept from 0 dBm to 25 dBm in order to find the optimum LO power level to maximize conversion gain. Fig. 5.4 shows the conversion gain versus LO power levels. It can be seen we have a higher conversion gain at 19 dBm of LO power; but this power level may not be a safe power level for the FETs as it is on the edge of the maximum input power rating of 16 dBm.

Fig. 5.4: Input LO power sweep analysis

Methods to Improve RF-IF Isolation

It was noticed from Fig. 5.3 that the RF-IF isolation is found to be very poor. For the retrodirective action to take place, this RF-IF isolation must be improved. Since, the IF frequency is very close to both the LO and RF frequencies it makes it difficult to have a good isolation. The LO-IF isolation has been taken care of by the virtue of the rat race coupler used in the input circuit. The following are some ideas to improve the RF-IF isolation.

Bandstop or Notch Filter:

This method works only if there is some frequency offset between the RF and IF (a minimum shift of 20 MHz have been used by researchers in the past). A possible notch filter uses transmission lines or cavity resonators. It uses either open-circuited or short-circuited filter to implement the bandpass or bandstop filters. The stubs and the transmission line sections are quarter wavelength at the center frequency. These filters provide very narrow bandwidths. Hence, can be an alternative to block the RF frequency which is slightly offset from the LO and IF. But, the disadvantage of these filters is they use very high impedance stubs, manufacturing of which can be difficult.

The idea here is to block the RF frequency. This can be done using a bandstop filter with a center frequency near the RF and having a narrow bandwidth. Using the filter design formulae given in Microwave engineering by David Pozar [18], the values of the stub impedances can be calculated. The values of the low pass prototype are provided in the appendix III.

Let the center frequency of the filter be 3.15 GHz. The RF frequency required to be stopped is 3.15 GHz. The impedances of the shunt stubs can be calculated using the formula:

$$Z_{on} = (4*Z_0/\pi*g_n*\Delta) \quad \text{Eq. 5.1}$$

Z_{on} are the impedances of the shunt stubs, Z_0 is the impedance of the filter (here it is taken as 50 ohm), g_n are the low pass prototype values calculated from appendix III, and Δ is the fractional bandwidth. With a chosen bandwidth of 10%, the values of Z_{on} are calculated.

Table 5.1: Calculated values of shunt stubs impedances

n	g_n	Z_{on}
1	1.5963	399
2	1.0967	584
3	1.5963	399

The filter was designed in Serenade SV and a linear analysis was performed from frequencies 2 GHz to 4 GHz. Fig. 5.5 shows the coupled resonator filter.

Fig. 5.5: Notch filter

A linear analysis is performed on the filter. Fig. 5.6 shows the filter response. It is seen that the attenuation at RF frequency is 130 dB. The attenuation at IF frequency is 10.5 dB is found. Hence, an isolation of more than 30 dB can be obtained between RF and IF with the use of these filters.

Fig. 5.6: Response of a coupled resonator filter

RF Suppression Using Out-of-Phase Fed RF to FETs:

The LO was fed out-of-phase to the FETs to provide better LO-IF isolation. Similarly, if the RF is fed out of phase to the FETs, a better RF isolation can be achieved. Fig. 5.7 shows the schematic to achieve better RF isolation. The RF is fed from the difference port of the hybrid II. Hence the RF signals to the hybrid III and IV are out of phase. The resulting RF at FET2 is 180-deg out-of-phase from

the RF at FET1. Hence, the RF leakage at the output, being out-of-phase will cancel out providing a better RF-IF isolation.

Fig. 5.7: RF-IF isolation using out-of-phase fed RF

The disadvantage of this method is it uses four hybrid couplers, and hence the amount of circuitry has increased. This method may not be space effective with large amount of circuitry.

Due to the limitations of *Serenade SV*, limits the number of objects to be used to 25, the simulation of the above method was not performed.

Use of Two Subharmonic Mixers for Better RF-IF Isolation:

The poor RF-IF isolation at the output of the subharmonic mixer can be improved by using two balanced subharmonic mixers [11]. Brabetz et al. proposed the use of two harmonic mixers; with RF signals being fed out-of-phase

to the mixers. Their design of the subharmonic mixer used diodes, passive devices. However, the same idea can be applied to the active mixers using FETs. The RF signals to both mixers are 180° out-of-phase using transmission lines. In addition, a phase difference of 90° is applied to the LO signal fed to the two mixers. This kind of an arrangement results in self cancellation of the RF and LO signals, providing better isolation. Fig. 5.8 shows the schematic for this kind of an arrangement.

Fig. 5.8: RF-IF suppression configuration [11]

To facilitate the discussion of RF suppression at the output, Fig. 5.1 has been shown here again as Fig. 5.9.

Fig. 5.9: Subharmonic Mixer Schematic

The LO and RF signals applied at the sum and difference ports of the balanced harmonic mixers are $V_{LO}\cos(\omega t_{LO})$ and $V_{RF}\cos(\omega t_{RF}+\theta)$, where θ is the phase term to be conjugated. In the harmonic mixer 1, the output at FET1 will be

$$\text{LO: } \omega t_{LO} - 90^\circ, \text{ RF: } \omega t_{RF} + \theta - 90^\circ$$

And similarly, the signals at the output of FET2

$$\text{LO: } \omega t_{LO} - 270^\circ, \text{ RF: } \omega t_{RF} + \theta - 90^\circ$$

The fundamental difference product at FET1, will be

$$(\omega t_{LO} - 90^\circ) - (\omega t_{RF} + \theta - 90^\circ) = \omega t_{LO} - \omega t_{RF} - \theta \quad (5.1)$$

The output difference product of the second harmonic of LO at FET1, ωt_D will be

$$2(\omega t_{LO} - 90^\circ) - (\omega t_{RF} + \theta - 90^\circ) = \omega t_D - \theta - 90^\circ$$

Where

$$\omega t_{IF} = \omega t_D = 2\omega t_{LO} - \omega t_{RF} \quad (5.2)$$

The fundamental difference product at FET2, will be

$$(\omega t_{LO} - 270^\circ) - (\omega t_{RF} + \theta - 90^\circ) = \omega t_{LO} - \omega t_{RF} - \theta - 180^\circ \quad (5.3)$$

Hence, from (5.1) and (5.3), the fundamental mixing component is cancelled as they are 180° out-of-phase.

The output difference product of the second harmonic of LO at FET2, ωt_D will be

$$2(\omega t_{LO} - 270^\circ) - (\omega t_{RF} + \theta - 90^\circ) = \omega t_D - \theta - 360^\circ - 90^\circ = \omega t_D - \theta - 90^\circ \quad (5.4)$$

the mixer outputs (5.2) and (5.4), add in phase giving the composite output as $\omega t_D - \theta - 90^\circ$

Now,

In the harmonic mixer 2, the output at FET1 will be

$$\text{LO: } \omega t_{LO} - 180^\circ, \text{ RF: } \omega t_{RF} + \theta - 270^\circ$$

And similarly, the signals at the output of FET2

LO: $\omega_{tLO} - 360^\circ$, RF: $\omega_{tRF} + \theta - 270^\circ$

The fundamental mixing component is cancelled the same way as it was shown for harmonic mixer 1.

The output difference product of the second harmonic of LO at FET1, ω_{tD} will be

$$2(\omega_{tLO} - 180^\circ) - (\omega_{tRF} + \theta - 270^\circ) = \omega_{tD} - \theta - 90^\circ$$

Where

$$\omega_{tD} = 2\omega_{tLO} - \omega_{tRF} \quad (5.5)$$

The output difference product of the second harmonic of LO at FET2, ω_{tD} will be

$$2(\omega_{tLO} - 360^\circ) - (\omega_{tRF} + \theta - 270^\circ) = \omega_{tD} - \theta - 450^\circ = \omega_{tD} - \theta - 90^\circ \quad (5.6)$$

the mixer outputs (5.5) and (5.6), add in phase giving the composite output as $\omega_{tD} - \theta - 90^\circ$

Finally, the harmonic products from the harmonic mixers 1 and 2 in Fig. 5.15 are added in phase using a power combiner in order to give maximum signal strength for the conjugated IF signals at the output port. The RF leakages from both mixers are $\omega_{tRF} + \theta - 90^\circ$ and $\omega_{tRF} + \theta - 270^\circ$, which are 180° out of phase hence, get cancelled. A good RF-IF isolation is achieved.

Experimental Design

The above subharmonic mixer has been simulated using ideal transmission lines (because *Serenade SV* does not allow the use CPW). But, as the thesis deals with implementing using CPW transmission lines, the individual components in the mixer design were simulated in *Ansoft Ensemble*. The subsequent sections details the design and simulation results of the coupler, dc bias for the transistors and the dc block.

Rat Race Coupler

A rat race coupler is one of the simplest 180-deg coupler to design [4]. A 180-deg hybrid or coupler is a 4 port device which provides two (C&D) equal

amplitude in phase signals when fed from its sum port, B (Σ), and two (C&D) equal amplitude 180 degrees out of phase signals when fed from its difference port, A (Δ). Fig. 5.10 shows the operation of a 180-deg hybrid.

Fig. 5.10: Operation of 180-deg hybrid [4]

Most of the ring hybrids use microstrip or striplines. Hence, a ring hybrid was designed in *Ansoft Ensemble*, with coplanar waveguides and slots. The port impedances are optimized using Ensemble estimator to 50 ohm. And the width of the slot (ring) is determined to be 0.48 mm for an impedance of 70.7 ohm. A full-wave analysis is performed on the ring hybrid shown in Fig. 5.11. With Port 3 excited, the scattering matrix for the ring hybrid is noted. Fig. 5.12 shows the simulated scattering matrix results for the ring hybrid. As seen in Fig. 5.12, the isolation between the port 1 and port 3 is well below 30 dB in the band of interest of 3 – 3.5 GHz. The input power is split equally between the port 2 and port 4 in the band. The power levels being 3 dB less than the input power level. The return loss for the port 3 is found to be -45 dB. The bandwidth ($|S_{23}|, |S_{43}| = 3 \pm 1$ db) is found to be 17 %. When the bandwidth is considered with $|S_{11}|, |S_{22}| \leq -10$ db as the criteria, then we have a 22 % bandwidth.

Fig. 5.11: Ring Hybrid using CPWs and slots

Fig. 5.12: Scattering matrix for the ring

The experimental setup and the measured results of the rat race coupler are shown in Fig. 5.13 and Fig. 5.14.

(a)

(b)

Fig. 5.13: Photographs of (a) rat race coupler (b) experimental setup of the coupler

(a)

(b)

(c)

Fig. 5.14: Experimental results of rat race coupler (a) $|S_{11}|$ (b) $|S_{21}|$ (c) $|S_{22}|$

From the Fig. 5.14, it can be seen that a good return loss for the ports i.e. $|S_{11}|$, $|S_{22}|$ was found. Also the transmission coefficient, $|S_{21}|$, is 3dB. The other port is isolated from the input port. The phase differences were also found to be in agreement to the simulation results. A bandwidth of 25% was found with $|S_{11}|$, $|S_{22}| \leq -10\text{dB}$ as the criteria.

DC Bias

To bias the FETs, a bias structure using two quarter wavelength transmission lines at fundamental frequency with different characteristic impedance are connected in a way such that the DC biasing for the transistors does not affecting the AC performance of the circuit. The dc bias circuit designed in *Ansoft Ensemble* is shown in fig. 5.15. At point A the termination is an open circuit which is transformed to a near short circuit termination at point B with the help of the low impedance quarter wavelength transformer. Now at point B the near short circuit is transformed to an open circuit at the point C where the bias circuit meets the 50 ohm line with the help of the high impedance quarter wave transformer. Hence, very high input impedance is seen at point C for fundamental and all higher harmonics along with an open circuit characteristics at DC.

Fig. 5.15: Transistor dc bias circuit

A full-wave analysis was performed on the structure shown in Fig. 5.15. Fig. 5.16 shows the scattering matrix for the dc bias. The signal from the coupler (port 1) is sent to the gate of the transistor (port 2). The S_{21} is almost 0 dB in the required band of 3-3.5 GHz. Hence, there is no power coupled to the bias circuitry. The S_{11} and S_{22} are found to be -30 dB in the band.

Fig. 5.16: DC bias scattering matrix

DC Block

For the dc applied to bias the transistor not to enter the rest of the circuit, we need a high value capacitor. The high impedance of the capacitance blocks the DC while allowing the high frequency signals to pass. The capacitances can be realized using interdigitated transmission line structures. They provide the

necessary discontinuity for the dc to be blocked. Fig. 5.17 shows the dc block designed in *Ansoft Ensemble*. The width of the 50 ohm transmission line is optimized for three teeth to fit. A full-wave analysis was performed on the structure; results are shown in Fig. 5.18. The S_{21} is found to be 0 dB in the required band, allowing the signal to pass from port 1 to port 2.

Fig. 5.17: Interdigitated dc block

Fig. 5.18: Scattering matrix for interdigitated dc block

DC Block and DC Bias combined

The dc block and the dc bias circuit were combined together to check if they work properly when put together. Fig. 5.19 shows the dc block and dc bias together. A full-wave analysis on the circuit gave good results in accordance with the individual components. The scattering matrix for the structure is shown in Fig. 5.20.

Fig. 5.19: DC-block and DC-bias combined

Fig. 5.20: Scattering matrix for dc block and dc bias

Subharmonic Mixer Schematic in Ansoft Ensemble

The subharmonic mixer layout with all the individual components was drawn in *Ansoft ensemble*. The mixer is built on RT/Duroid substrate with $\epsilon_r = 10.2$ and thickness of 50 mil. Slots were left for the FETs to be soldered after it is built using *ISOPRO* and *T-Tech* milling machine. Fig. 5.21 shows the layout of the subharmonic mixer. Long high impedance lines are used to provide bias for the FETs to reduce the fabrication difficulties.

Fig. 5.21: Layout of the subharmonic mixer in *Ansoft Ensemble*

CHAPTER 6

CONCLUSIONS AND FUTURE WORK

Conclusions

The aim of this thesis was to design an active balanced subharmonic retrodirective antenna array. A novel dual-polarized antenna was modeled in the commercial software package *Ansoft Ensemble*. Two orthogonal conductor-backed coplanar waveguide feeds were used to achieve the dual polarization. Aperture coupling was used to couple the microwave energy to the radiating element. Full-wave analysis was performed and the results were presented. Very good return loss and isolation between ports were noted. Also, dual polarization was achieved.

The antenna design was built to test the simulation results. Much to our despair, due to primitive fabrication facilities the performance of the antenna was poor. It may have to do with the proper alignment of the apertures in the ground plane with the patch and the feeds. The air gap between the two substrates due to the accumulated solder may also cause the performance degradation. Simulation results with an additional air gap were also provided.

A balanced subharmonic mixer using FETs as mixing devices was designed. Simulations were performed in *Serenade SV*, as the active devices cannot be modeled in *Ansoft Ensemble*. Spectral analysis of the subharmonic mixer showed a very good LO-IF isolation and a conversion gain. The FETs effectively short-circuit each other at fundamental mixing frequency. A sweep analysis was performed to determine the optimum LO power to produce a good conversion gain. To improve the RF-IF isolation, three methods have been proposed. Further, as the design of the subharmonic mixer involved ideal transmission lines (due to the limitation of *Serenade SV*), each of the components

(rat race coupler, DC bias, DC Block) were designed separately in *Ansoft Ensemble*. Full-wave analysis was performed on these components and results were provided.

Suggestions for future work

The dual polarized antenna can be built and tested for better performance. Far-field patterns of the antennas should be measured.

Further work could include construction of the complete retrodirective antenna array and measurement of its performance. Unfortunately, such measurements are currently beyond the capabilities of the microwave lab facilities. A compact design for the array would be especially attractive. The realization of the RF-IF isolation methods could form a part of the future work.

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APPENDIX A

DATA SHEETS FOR NE722S01 (N - CHANNEL GaAs MESFET)

FEATURES

- **HIGH POWER GAIN:**
GS = 6 dB TYP at $f = 12$ GHz
- **OUTPUT POWER** (at 1 dB compression):
15 dB TYP at $f = 12$ GHz
- **LOW NOISE/HIGH GAIN:**
NF = 0.9 dB TYP, Ga = 12 dB TYP at $f = 4$ GHz
- **GATE LENGTH:** LG = 0.8 μ m (recessed gate)
- **GATE WIDTH:** WG = 400 μ m

DESCRIPTION

The NE722S01 is a low cost GaAs MESFET suitable for both amplifier and oscillator applications through X-band. The device features a 0.8 micron recessed gate, triple epitaxial technology and is fabricated using ion implantation for improved RF and DC performance and uniformity. This device's low phase noise and high f_T makes it an excellent choice for oscillator applications on a digital LNB (Low Noise Block). The NE722S01 is housed in a low cost plastic package which is available in Tape and Reel.

NEC's stringent quality assurance and test procedures ensure the highest reliability performance.

APPLICATIONS

- C to X band low noise amplifiers
- C to X band oscillators

OUTLINE DIMENSION (Units in mm)

PACKAGE OUTLINE S01

ELECTRICAL CHARACTERISTICS ($T_A = 25^\circ\text{C}$)

PART NUMBER PACKAGE OUTLINE			NE722S01 S01		
SYMBOLS	PARAMETERS AND CONDITIONS	UNITS	MIN	TYP	MAX
I_{GSS}	Gate to Source Leak Current, $V_{GS} = -5$ V	mA		1.0	10
I_{DSS}	Saturated Drain Current, $V_{DS} = 3$ V, $V_{GS} = 0$ V	mA	60	90	120
V_{GS}	Gate to Source Cutoff Voltage, $V_{DS} = 3$ V, $I_D = 100$ μ A	V	-0.5	-	-4.0
g_m	Transconductance, $V_{DS} = 3$ V, $I_{DS} = 30$ mA	mS	20	45	-
GS	Power Gain, $V_{DS} = 3$ V, $I_{DS} = 30$ mA, $f = 12$ GHz	dB		6	-
P1dB	Output Power at 1 dB Gain Compression Point at $V_{DS} = 3$ V, $I_{DS} = 30$ mA, $f = 12$ GHz	dBm		15.0	
NF	Noise Figure, $V_{DS} = 3$ V, $I_{DS} = 10$ mA, $f = 4$ GHz	dB		0.9	-
Ga	Associated Gain, $V_{DS} = 3$ V, $I_{DS} = 10$ mA, $f = 4$ GHz	dB		12	

NE722S01

ABSOLUTE MAXIMUM RATINGS¹ (TA = 25°C)

SYMBOLS	PARAMETERS	UNITS	RATINGS
VDS	Drain to Source Voltage	V	5.0
VGS	Gate to Source Voltage	V	-5.0
VGD	Gate to Drain Voltage	V	-6.0
ID	Drain Current	mA	loss
PT	Total Power Dissipation	mW	250
PIN	Input Power	mW	40
TCH	Channel Temperature	°C	125
TSTG	Storage Temperature	°C	-65 to +125

Note:

1. Operation in excess of any one of these parameters may result in permanent damage

TYPICAL PERFORMANCE CURVES (TA = 25°C)

Note:

1. Gain Calculation:

$$MAG = \frac{|S_{21}|}{|S_{12}|} \left(K \pm \sqrt{K^2 - 1} \right), \text{ When } K \leq 1, \text{ MAG is undefined and MSG values are used. } MSG = \frac{|S_{21}|}{|S_{12}|}, K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2 |S_{12}| |S_{21}|}, \Delta = S_{11} S_{22} - S_{21} S_{12}$$

MAG = Maximum Available Gain

MSG = Maximum Stable Gain

RECOMMENDED OPERATING CONDITIONS (TA = 25°C)

PART NUMBER			NE722S01		
SYMBOLS	PARAMETERS	UNITS	MIN	TYP	MAX
VDS	Drain to Source Voltage	V		3	4
ID	Drain Current	mA		30	40

ORDERING INFORMATION

PART NUMBER	SUPPLY FORM	MARKING
NE722S01-T1	Tape & Reel 1000 pos/reel	P
NE722S01-T1B1	Tape & Reel 4000 pos/reel	

Note:

1. Available if quantity is over 100k per month

TYPICAL PERFORMANCE CURVES ($T_A = 25^\circ\text{C}$)**OUTPUT POWER vs. INPUT POWER**

NE722S01

TYPICAL SCATTERING PARAMETERS (TA = 25°C)

NE722S01

V_{DS} = 3.0 V, I_{DS} = 10 mA

FREQUENCY GHz	S ₁₁		S ₂₁		S ₁₂		S ₂₂	
	MAG	ANG	MAG	ANG	MAG	ANG	MAG	ANG
2.0	0.912	-44.0	3.100	136.2	0.077	59.0	0.659	-30.2
2.5	0.876	-56.1	3.037	124.9	0.091	51.2	0.629	-38.0
3.0	0.828	-68.0	2.935	113.9	0.105	42.9	0.597	-46.4
3.5	0.784	-79.3	2.819	103.7	0.115	36.2	0.570	-53.5
4.0	0.737	-89.5	2.696	94.2	0.124	30.0	0.546	-60.9
4.5	0.699	-99.3	2.590	85.3	0.130	24.7	0.529	-67.2
5.0	0.660	-109.0	2.499	76.6	0.136	19.1	0.514	-73.6
5.5	0.620	-119.0	2.420	67.9	0.140	14.1	0.495	-79.6
6.0	0.583	-130.6	2.355	59.4	0.146	9.1	0.475	-85.2
6.5	0.547	-143.8	2.283	50.1	0.148	3.8	0.447	-91.5
7.0	0.516	-158.5	2.196	41.0	0.151	-1.6	0.408	-97.0
7.5	0.490	-173.7	2.098	32.2	0.149	-6.6	0.366	-103.7
8.0	0.500	172.6	2.016	23.7	0.152	-10.1	0.331	-110.8
8.5	0.510	159.9	1.920	15.3	0.150	-13.6	0.298	-120.4
9.0	0.526	148.4	1.834	7.2	0.151	-17.3	0.274	-133.2
9.5	0.540	138.4	1.749	-0.6	0.151	-20.3	0.265	-147.9
10.0	0.553	129.9	1.676	-7.9	0.152	-23.2	0.275	-160.7
10.5	0.566	120.6	1.608	-15.8	0.156	-26.3	0.297	-172.7
11.0	0.576	111.3	1.542	-23.5	0.157	-29.8	0.312	-178.5
11.5	0.592	101.8	1.470	-31.1	0.157	-32.9	0.328	-171.0
12.0	0.608	92.8	1.401	-38.4	0.158	-35.7	0.340	-163.5
12.5	0.640	85.2	1.325	-45.7	0.159	-38.0	0.339	-155.6
13.0	0.665	79.1	1.256	-52.7	0.160	-41.3	0.341	-145.5
13.5	0.693	73.3	1.183	-59.7	0.158	-44.5	0.356	-133.8
14.0	0.716	69.3	1.111	-66.4	0.162	-47.0	0.366	-122.3
14.5	0.744	64.8	1.045	-73.3	0.163	-49.5	0.421	-111.6
15.0	0.759	59.6	0.966	-79.7	0.159	-53.4	0.474	-104.4
15.5	0.756	55.5	0.893	-85.4	0.159	-56.8	0.516	-98.7
16.0	0.750	51.0	0.839	-91.0	0.160	-57.6	0.563	-95.8
16.5	0.738	45.4	0.777	-96.7	0.158	-61.4	0.601	-92.8
17.0	0.728	40.9	0.714	-101.5	0.158	-63.1	0.624	-89.4
17.5	0.721	36.4	0.676	-105.7	0.156	-65.9	0.628	-86.9
18.0	0.721	32.5	0.624	-109.8	0.158	-68.4	0.625	-82.2

Note

1. Gain Calculation:

$$\text{MAG} = \frac{|S_{21}|}{|S_{12}|} \left(K \pm \sqrt{K^2 - 1} \right), \text{ When } K \leq 1, \text{ MAG is undefined and MSG values are used. } \text{MSG} = \frac{|S_{21}|}{|S_{12}|}, K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2 |S_{12}| |S_{21}|}, \Delta = S_{11} S_{22} - S_{21} S_{12}$$

MAG = Maximum Available Gain

MSG = Maximum Stable Gain

TYPICAL SCATTERING PARAMETERS (TA = 25°C)

NE722S01

V_{DS} = 3.0 V, I_{DS} = 30 mA

FREQUENCY GHz	S ₁₁		S ₂₁		S ₁₂		S ₂₂	
	MAG	ANG	MAG	ANG	MAG	ANG	MAG	ANG
2.0	0.896	-48.3	3.721	134.0	0.083	59.9	0.547	-29.6
2.5	0.851	-61.5	3.606	122.4	0.077	53.2	0.519	-37.4
3.0	0.799	-74.1	3.446	111.3	0.086	45.1	0.489	-45.4
3.5	0.753	-86.1	3.275	101.1	0.095	39.6	0.464	-52.6
4.0	0.705	-97.0	3.102	91.6	0.102	34.1	0.443	-59.5
4.5	0.666	-107.3	2.957	82.7	0.106	29.5	0.431	-65.5
5.0	0.626	-117.5	2.834	74.0	0.111	24.9	0.419	-71.5
5.5	0.586	-128.3	2.724	65.3	0.115	20.6	0.406	-76.9
6.0	0.553	-140.6	2.627	56.8	0.122	17.0	0.396	-82.2
6.5	0.521	-154.2	2.522	47.7	0.125	12.6	0.386	-88.2
7.0	0.497	-169.4	2.402	39.0	0.128	8.5	0.334	-93.4
7.5	0.489	175.8	2.281	30.5	0.128	5.0	0.299	-99.7
8.0	0.501	162.9	2.176	22.5	0.131	1.8	0.267	-106.8
8.5	0.519	151.0	2.062	14.5	0.135	-1.2	0.234	-117.8
9.0	0.540	140.4	1.963	6.8	0.139	-3.4	0.209	-132.7
9.5	0.562	131.2	1.865	-0.7	0.142	-6.8	0.206	-150.0
10.0	0.575	123.2	1.786	-7.8	0.146	-9.9	0.221	-164.3
10.5	0.589	114.3	1.709	-15.4	0.151	-12.8	0.246	-177.2
11.0	0.602	105.5	1.637	-22.9	0.156	-16.7	0.265	173.7
11.5	0.619	96.4	1.554	-30.2	0.161	-20.5	0.285	166.1
12.0	0.633	88.0	1.480	-37.4	0.163	-23.9	0.299	158.1
12.5	0.666	80.9	1.400	-44.5	0.168	-27.4	0.303	149.9
13.0	0.690	75.2	1.321	-51.1	0.171	-30.5	0.307	139.1
13.5	0.715	70.0	1.241	-57.9	0.173	-33.9	0.326	127.4
14.0	0.740	66.4	1.169	-64.2	0.173	-37.3	0.356	116.2
14.5	0.769	62.0	1.101	-70.8	0.174	-41.0	0.396	106.2
15.0	0.780	57.0	1.021	-77.1	0.178	-45.3	0.450	99.5
15.5	0.778	52.5	0.945	-82.6	0.175	-48.6	0.494	94.8
16.0	0.774	48.2	0.886	-88.0	0.176	-51.4	0.540	92.0
16.5	0.759	42.7	0.824	-93.4	0.176	-56.1	0.576	89.4
17.0	0.750	38.1	0.767	-98.3	0.174	-58.3	0.601	89.3
17.5	0.739	33.9	0.721	-102.3	0.175	-61.1	0.604	84.0
18.0	0.741	30.2	0.672	-106.6	0.176	-63.3	0.599	79.2

Note

1. Gain Calculation:

$$MAG = \frac{|S_{21}|}{|S_{12}|} \left(K \pm \sqrt{K^2 - 1} \right), \text{ When } K \leq 1 \text{ MAG is undefined and MSG values are used. } MSG = \frac{|S_{21}|}{|S_{12}|} \quad K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2 |S_{12}| S_{21}}, \Delta = S_{11} S_{22} - S_{21} S_{12}$$

MAG = Maximum Available Gain

MSG = Maximum Stable Gain

NONLINEAR MODEL

SCHEMATIC

FET NONLINEAR MODEL PARAMETERS (1)

Parameters	Q1	Parameters	Q1
VTO	-2.24	RG	8
VTOSC	0	RD	0.5
ALPHA	8	RS	5
BETA	0.055	RCMET	0
TGAMMA	0.04	TNOM	27
TGAMMAAC	0.05	XTI	3
Q	1.5	EG	1.43
TODELTA	0.25	VTOTC	0
VBI	1	BETATUE	0
IS	1e-14	FFE	1
N	1.3	FNCP	150e-6
RIS	0	R	0.5
RID	0	P	1
TAU	3e-12	C	0.9
CDS	0.19e-12		
RDB	250		
CBS	1e-9		
CGS	0.42e-12		
CGD	0.05e-12		
DELTA1	0.3		
DELTA2	0.3		
FC	0.5		
VBR	Infinity		

(1) ADS TOM Model

(2) To simulate phase noise using AF/KF: AF = 1.5 KF = 2e-10

UNITS

Parameter	Units
time	seconds
capacitance	farads
inductance	henries
resistance	ohms
voltage	volts
current	amps

MODEL RANGE

Frequency: 0.1 to 18 GHz

Bias: Vbs = 2 V to 4 V, Id = 10 mA to 40 mA

Date: 02/2002

Life Support Applications

These NEC products are not intended for use in life support devices, appliances, or systems where the malfunction of these products can reasonably be expected to result in personal injury. The customers of CEL using or selling these products for use in such applications do so at their own risk and agree to fully indemnify CEL for all damages resulting from such improper use or sale.

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02/13/2002

APPENDIX B

TOM MODEL FOR MESFETs

Figure 1: TOM-2 Model.

1 Introduction

TOM-2 is an improved model for GaAs MESFETs. It is based on the original Triquint model (aka. TOM, herein referred to as TOM-1) with refinements to improve accuracy in the knee and subthreshold region. Particular attention was given to the temperature effects.

The model has been implemented using Pspice version 6.1 but should be readily portable to other analog simulators.

This model does not claim to be the final word in MESFET models—it is simply a next order effort to improve the overall accuracy of the model based on several years of experience with TOM-1. TOM-2 retains the desirable

features of TOM-1 while improving performance in the subthreshold (V_{gs}) near cut-off, and knee regions (V_{ds} of 1 volt or less). Additional temperature coefficients are included related to the drain current, and major deficiencies in the behavior of the capacitance as a function of temperature are corrected.

2 Equivalent Circuit

Figure 1 shows the basic model used in TOM-2. The model details specify how the current sources and the variable capacitors depend on bias conditions and on temperature.

The equations given in the rest of this document reference the voltages across the "intrinsic" portion of the FET as outlined in the figure. These intrinsic voltages are denoted as V_{gs} and V_{ds} for the gate-to-source and drain-to-source respectively.

Note that the model does not provide a mechanism for modeling the low frequency drain conductance dispersion that is well known in GaAs MESFETs. We model this effect with an external subcircuit with feedback coupled through a low pass RC network (not described in this document).

3 Basic Equations

3.1 Current source equations

The following expression is used to compute I_{ds} for the current source in figure 1.

$$I_{ds} = \frac{I_{dso}}{1 + \delta V_{ds} I_{dso}} \quad (1)$$

Where:

$$I_{dso} = w \cdot B_g^{-Q} \cdot F_0(x) V_{ds} \quad (2)$$

$$F_0(x) = \frac{x}{\sqrt{1+x^2}} \quad (3)$$

and V_g is given by:

$$V_g = QV_{st} \ln \left[\exp \left(\frac{V_{gs} - V_{to} + \gamma V_{ds}}{QV_{st}} \right) + 1 \right] \quad (4)$$

where V_{to} and γ are constants. w is the device width.

In general, we expect V_{st} to be near the thermal voltage kT/q , or about 0.026 mV at 25°C. This leads us to present the V_{st} by an ideality factor, N_{st} thus:

$$V_{st} = N_{st} \left(\frac{kT}{q} \right) \quad (5)$$

In addition, it is observed that the subthreshold slope varies slightly with V_{ds} , so we put in a linear dependence to account for this:

$$N_{st} = N_g + N_d V_{ds} \quad (6)$$

Notice that $N_{st} = 0$ corresponds to no subthreshold—that is, the current cuts off sharply at threshold. This behavior is similar to a diode where setting the ideality factor equal to zero results in a piecewise linear behavior.

The gate current, I_g is the sum of the current flowing in the gate-drain diode, I_{ds} , and the gate-source diode, I_{gs} . These currents are given by the classical diode formula:

$$I_g = I_{gs} + I_{gd} \quad (7)$$

$$I_{gs} = wI_s(\epsilon^{qV_{gs}/nkT} - 1) \quad (8)$$

$$I_{gd} = wI_s(\epsilon^{q(V_{gs}-V_{ds})/nkT} - 1) \quad (9)$$

Notice also that the total drain current, I_d includes a component flowing from the gate:

$$I_d = I_{ds} - I_{gd} \quad (10)$$

Usually, the above equations will be incorporated into a subroutine that must deliver the currents I_d and I_g back to the simulation engine.

5 Model Parameters and Scaling

Parameter	Symbol	Description	Units	Default	Scaling
LEVEL	—	model index (-1 for TQM-2)			
VTO	V_t	Threshold voltage	volt	-2.5	—
ALPHA	α	Knee-voltage parameter	volt ⁻¹	2.0	—
BETA	β	Transconductance parameter	amp/volt ⁻²	0.1	—
GAMMA	γ	Threshold shifting parameter	volt ⁻¹	0	—
DELTA	δ	Output feedback parameter	volt	0.2	1/ w
Q	Q	Power-law parameter	—	2	—
NG	N_g	Subthreshold slope gate parameter	—	0	—
ND	N_d	Subthreshold slope drain pull parameter	volt ⁻¹	0	—
TAU	τ	Conduction current delay time	sec	0	—
RG	R_g	Gate ohmic resistance	ohm	0	—
RD	R_d	Drain ohmic resistance	ohm	0	1/ w
RS	R_s	Source ohmic resistance	ohm	0	1/ w
IS	I_s	Gate diode saturation current	amp	1×10^{-14}	—
N	n	Gate diode ideality factor	—	1	—
VBI	V_{bi}	Gate diode built-in potential	volt	1.0	—
VDELTA	V_δ	Capacitance transition voltage	volt	0.2	—
VMAX	V_{max}	Gate diode capacitance limiting voltage	volt	0.05	—
CGD	C_{gd}	Gate-to-drain "zero-bias" capacitance	farad	0	w
CGS	C_{gs}	Gate-to-source "zero-bias" capacitance	farad	0	w
CDS	C_{ds}	Drain-to-source capacitance	farad	0	w
EG	E_g	Barrier height	volt	1.11	—
XTI	X_{ti}	I_s temperature exponent	—	0	—
VTOTC	V_{to}^t	V_{to} temperature coefficient (linear)	volt/ $^{\circ}\text{C}$	0	—
VBITC	V_{bi}^t	V_{bi} temperature coefficient (linear)	volt/ $^{\circ}\text{C}$	0	—
ALPHATCE	α^t	α temperature coefficient (exponential)	%/ $^{\circ}\text{C}$	0	—
BETATCE	β^t	β temperature coefficient (exponential)	%/ $^{\circ}\text{C}$	0	—
GAMMATC	γ^t	Linear temperature coefficient for γ	$^{\circ}\text{C}^{-1}$	0	—
TRG1	R_g^t	Linear temperature coefficient for R_g	$^{\circ}\text{C}^{-1}$	0	—
TRD1	R_d^t	Linear temperature coefficient for R_d	$^{\circ}\text{C}^{-1}$	0	—
TRS1	R_s^t	Linear temperature coefficient for R_s	$^{\circ}\text{C}^{-1}$	0	—
CGDTCE	C_{gd}^t	Linear temperature coefficient for C_{gd}	$^{\circ}\text{C}^{-1}$	0	—
CGSTCE	C_{gs}^t	Linear temperature coefficient for C_{gs}	$^{\circ}\text{C}^{-1}$	0	—
KF	K_f	Flicker noise coefficient		0	—
AF	A_f	Flicker noise exponent		1	—

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