

**T.C.
AKDENİZ ÜNİVERSİTESİ**



**YAMA GİYDİRİLMİŞ DALGA KILAVUZU (YGDK)
TEKNOLOJİSİ İLE X BANDI
BANT GEÇİREN FİLTRE TASARIMI**

Mohammed R. A. NASSER

**FEN BİLİMLERİ ENSTİTÜSÜ
ELEKTRİK ELEKTRONİK MÜHENDİSLİĞİ
ANABİLİM DALI
YÜKSEK LİSANS TEZİ**

HAZİRAN 2020

ANTALYA

**T.C.
AKDENİZ ÜNİVERSİTESİ**



**YAMA GİYDİRİLMİŞ DALGA KILAVUZU (YGDK)
TEKNOLOJİSİ İLE X BANDI
BANT GEÇİREN FİLTRE TASARIMI**

Mohammed R. A. NASSER

**FEN BİLİMLERİ ENSTİTÜSÜ
ELEKTRİK ELEKTRONİK MÜHENDİSLİĞİ
ANABİLİM DALI
YÜKSEK LİSANS TEZİ**

HAZİRAN 2020

ANTALYA

**T.C.
AKDENİZ ÜNİVERSİTESİ
FEN BİLİMLERİ ENSTİTÜSÜ**

**YAMA GİYDİRİLMİŞ DALGA KILAVUZU (YGDK)
TEKNOLOJİSİ İLE X BANDI
BANT GEÇİREN FİLTRE TASARIMI**

**Mohammed R. A. NASSER
ELEKTRİK ELEKTRONİK MÜHENDİSLİĞİ
ANABİLİM DALI
YÜKSEK LİSANS TEZİ**

**Bu tez Akdeniz Üniveristesi Bilimsel Araştırma Projeleri Koordinasyon Birimi
tarafından FYL-2020-5103 nolu proje ile desteklenmiştir.**

HAZİRAN 2020

T.C.
AKDENİZ ÜNİVERSİTESİ
FEN BİLİMLERİ ENSTİTÜSÜ

YAMA GİYDİRİLMİŞ DALGA KILAVUZU (YGDK)
TEKNOLOJİSİ İLE X BANDI
BANT GEÇİREN FİLTRE TASARIMI

Mohammed R. A. NASSER
FEN BİLİMLERİ ENSTİTÜSÜ
ANABİLİM DALI
YÜKSEK LİSANS TEZİ

Bu tez 19/Haziran/2020 tarihinde jüri tarafından Oybirliği / Oyçokluğu ile kabul edilmiştir.

Prof. Dr. Selcuk HELHEL (Danışman)

Prof. Dr. Şükrü ÖZEN

Dr. Öğretim Üyesi İ. Bahadır BAŞYİĞİT

ÖZET

YAMA GIYDİRİLMİŞ DALGA KILAVUZU (YGDK) TEKNOLOJİSİ İLE X BANDI BANT GEÇİREN FİLTRE TASARIMI

Mohammed R. A. NASSER

Yüksek Lisans Tezi, Elektrik Elektronik Mühendisliği Anabilim Dalı

Danışman: Prof. Dr. Selçuk HELHEL

Haziran 2020; 39 sayfa

Düzlemsel bant geçiren filtreler temel olarak herhangi bir RF / Mikrodalga cihazının önemli bir parçası olarak kabul edilir. İletişim ve radar sistemlerine yönelik artan talepler, daha yekpare ve hafif yapılardan üstün performans sağlayan ve diğer devrelerle kolayca entegre edilebilen filtreler gerektirir. Bu tezde, metal dalga kılavuzları ile mikroşerit yapıların özelliklerini birlikte taşıyan Yama Giydirilmiş Dalga Kılavuzu (YGDK) adı verilen yeni bir düzlemsel teknoloji ele alınmaktadır.

Bu tezde Yama Giydirilmiş Dalga Kılavuzu teknolojisini ve Çentikli Taban Yapılar (ÇTY) kullanarak iki adet üçüncü dereceden yekpare filtre tasarladık. Her iki tasarım da mikroşerit teknolojisiyle geniş bant adaptasyonu sağlamak için YGDK-Micro şerit konik geçişi kullanılmıştır.

Çalışmada X-bandı uygulaması için tasarlanmış bir Geniş Bant bant geçiren filtre tasarlanmıştır. Filtre, YGDK ve YGDK'nın hem üst hem de alt düzlemine yüklenen iki tip Çentikli Taban Yapılar (ÇTY) tarafından oluşturulmuş olup 10 GHz'de kesirli bant genişliği (FBW = 40%) şeklindedir. Üretim kolaylıkları nedeni ile ayrıca C-bandı bölgesinde çalışan 6.175 GHz' merkez frekansında 500 MHz bant genişliğine sahip ilk huzme genişliği (FBW = 8.3%) olan bir filtredir. YGDK'nın üst düzleminde YGDK ve 3-ÇTY hücreleri aracılığı ile oluşturulmuştur.

ANAHTAR KELİMELELER: Bant geçiren filtreler, Yama Giydirilmiş Dalga Kılavuzu (YGDK), Çentikli Taban Yapılar (ÇTY), Geniş Bant Konik geçiş, X-bandı, C-bandı.

JÜRİ:

Prof. Dr. Selcuk HELHEL (Danışman)

Prof. Dr. Şükrü ÖZEN

Dr. Öğretim Üyesi İ. Bahadır BAŞYİĞİT

ABSTRACT

FILTER DESIGN USING THE SUBSTRATE INTEGRATED WAVEGUIDE (YGDK) TECHNOLOGY ON X BAND

Mohammed R. A. NASSER

MSc Thesis in Electrical & Electronics Engineering

Supervisor: Prof. Dr. SELÇUK HELHEL

June 2020; 39 pages

Planer band-pass filters are considered as a major part of any RF/Microwave device. The increasing demands on communications and radar systems require filters that provide all of compactness and lightweight with superior performance and easy to integrate with other circuits. In this thesis, we look at new planer technology called substrate integrated waveguides (YGDK), developed recently with characteristics mixes the affirmative characteristics of the metallic waveguides and microstrip technologies, has opened the horizons in microwave filter design to provide new designs to match the high requirements of the increased demand. In this thesis, we have designed two 3rd order compact filters using the Substrate integrated waveguide (YGDK) technology and the Defected ground structure (ÇTY). Both of the designs are using the YGDK -microstrip taper transition to ensure wideband adaption with microstrip technology.

The first filter is a Wideband bandpass filter designed for X-band application has a center frequency at 10 GHz, with fractional bandwidth (FBW 40%), the structure composed by YGDK and two types of defected ground structures (ÇTY) loaded in both of the Top and bottom plane of the YGDK and its characterized by the wide free of spurious upper stopband, while the second filter is a broadband filter designed for C-band Space communications has a center frequency at 6.175 GHz, with 500 MHz bandwidth (FBW=8.3%), the structure composed by YGDK and 3-ÇTY cells on the top plane of the YGDK. The filters were built on a single-layer Rogers DiClad 880 substrate, and they were designed and optimized using the time domain solver "the Finite-difference time-domain method (FDTD) numerical method" in CST MW studio.

KEYWORDS: Band-pass filters, substrate-integrated waveguide (YGDK), defected ground structures (ÇTY), wideband taper transition, X-band, C-band.

COMMITTEE:

Prof. Dr. Selcuk HELHEL (Supervisor)

Prof. Dr. Şükrü ÖZEN

Asistant Prof. Dr. İ. Bahadır BAŞYİĞİT

ÖNSÖZ

Yekpare boyut, düşük insersiyon kaybı, iyi geri dönüş kaybı ve pürüzsüz grup gecikmesi. Bu filtrelerle ilgili tüm gereksinimler, geleneksel teknolojiler tarafından bir anda karşılanamaz, çünkü yüksek performans her zaman hantal bir boyuta sahip düzlemsel dışı dalga kılavuzu teknolojisi ile ilgilidir. Düzlemsel teknolojilerine gelince, diğer devrelerle yekparelik ve entegrasyon kolaylığı özellikleriyle birlikte bu profesyoneller kırılgen performans pahasına gelir. Bu tezde Yama Giydirilmiş Dalga Kılavuzu (YGDK) adı verilen ve düzlemsel teknolojilerinin kırılgen performansının ortadan kaldırılmasına yardımcı olan ve filtre tasarımı konusunda yeni ilkeler sağlayan yeni teknolojiyi tartışıyoruz.

Beni öğrencisi olarak kabul ettiği ve araştırmamı desteklediği için danışmanım Prof. Dr. Selçuk HELHEL'e içten şükranlarımı sunuyorum ve ölçümlerde yardımcı olduğu için Prof. Bektaş ÇOLAK'a teşekkür ederim.

Bana yardım etmeksizin gerçeğe dönüşmeyecek hayalimi gerçekleştirmeme yol açan, bana mali destek verdiği için Türk bursuna (YTB) en içten teşekkürlerimi sunuyorum. Ayrıca araştırmamı desteklediği için Akdeniz Üniversitesi Bilimsel Proje Destek Birimi'ne (BAP) teşekkür ederim. Ayrıca, bu çalışmada kullanılan tüm tesislerin Devlet Planlama Teşkilatı- Türkiye tarafından verildiği için Akdeniz Üniversitesi EMUMAM müdürlüğüne teşekkür etmek istiyorum (Hibe Numarası: 2007K120530-DPT).

Son olarak, paha biçilmez desteği ve sürekli teşvikleri için aileme, özellikle anneme ve arkadaşlarıma teşekkür ediyorum.

İÇİNDEKİLER

ÖZET.....	i
ABSTRACT.....	ii
ÖNSÖZ	iii
AKADEMİK BEYAN	v
SİMGELER VE KISALTMALAR.....	vi
ŞEKİLLER DİZİNİ.....	ix
1. GİRİŞ	1
2. KAYNAK TARAMASI	3
2.1. İletim Hatları.....	3
2.1.1. Mikroşerit Hatları (HONG, 2011)	3
2.1.2. Dikdörtgen Dalga Kılavuzları.....	4
2.1.3. Yama Giydirilmiş Dalga Kılavuzu (YGDK)	7
2.2. Mikroşerit – YGDK geçişi	11
2.3. Çentikli Taban Yapılar (ÇTY)	13
3. YÖNTEM.....	14
4. TASARIM VE BENZETTİM SONUÇLARI	16
4.1. X– Bant Filtresi	16
4.2. C– Bant Filtresi.....	19
5. ÖLÇÜMLER.....	22
5.1. X-Bant ölçümleri	24
6. SONUÇLAR	36
7. KAYNAKLAR	37
8. EKLER.....	39
8.1. EK-1 Tezin İngilizce versiyonu	39
ÖZGEÇMİŞ	

AKADEMİK BEYAN

Yüksek Lisans Tezi olarak sunduğum “**Yama Giydirilmiş Dalga Kılavuzu (YGDK) Teknolojisi ile X Bandı Bant Geçiren Filtre Tasarımı**” adlı bu çalışmanın, akademik kurallar ve etik değerlere uygun olarak yazıldığını belirtir, bu tez çalışmasında bana ait olmayan tüm bilgilerin kaynağını gösterdiğimi beyan ederim.

19/06/2020

Mohammed R. A. NASSER

SİMGELER VE KISALTMALAR

Simgeler

Hz	: Hertz (1/ saniye)
MHz	: Mega (10 ⁶) hertz
GHz	: Giga (10 ⁹) hertz
ϵ_r	: Göreli geçirgenlik veya dielektrik sabiti
$\tan \delta$	: Dielektrik kayıp
W	: Mikroşerit iletim hattı genişliği
T	: Plaka kaplama kalınlığı
h	: Plaka kalınlığı
μ	: Geçirgenlik
ϵ_{re}	: Etkili dielektrik sabiti
Z_c	: Karakteristik empedans
mm	: Milimetre
mm ²	: Kare milimetre
dB	: Desibel
ns	: Nanosaniye

Kısaltmalar

RF : Radyo frekansı

YGDK: Yama Giydirilmiş Dalga Kılavuzu

ÇTY : Çentikli Taban Yapılar

RWG : Dikdörtgen dalga kılavuzu

BDK : Baskılı devre kartı

FBW : Kesirli bant genişliği

X-bandı : 8.0 ile 12.0 gigahertz arasında değişen frekanslar

C-bandı : 4.0 ile 8.0 gigahertz arasında değişen frekanslar

YTB : Yurtdışı Türkler ve Akraba Topluluklar Başkanlığı

EM : Elektromanyetik

TL : İletim hattı

TEM : Enine elektromanyetik

DC : Doğru akım

F : Frekans

F_c : Kesim frekansı

TE_{10} : Baskın mod

λ_c : Kesme frekansına karşılık gelen dalga boyu

a_{eff} : YGDK genişliği, dikdörtgen dalga kılavuzunun genişliğine eşittir

C_0 : Boş alanda dalga hızı

CPW : Yarı-düzlemsel dalga kılavuzu

PTFE : Politetrafloroetilen

CST : Elektromanyetik benzetim yazılımı

MD : Mikrodalga

S -parametreliler : Saçılma parametreleri

S_{21} : İleri iletim (port 1'den port 2'ye)

S_{11} : Eşleşen bir yük tarafından sonlandırılan ağ çıkışı ile giriş yansıma katsayısı



ŞEKİLLER DİZİNİ

Şekil 1. 1. Basitleştirilmiş RF ön uç alıcı-verici blok şeması	1
Şekil 1. 2. Frekans seçiciliğine göre filtre türleri	1
Şekil 1. 3. Yama giydirilmiş dalga kılavuzunun (YGDK) genel yapısı.	2
Şekil 2. 1. Mikroşerit çizgi genel yapısı	3
Şekil 2. 2. Kesikli çizgilerin manyetik alan çizgilerini ve düz çizgilerin elektrik alan çizgilerini temsil ettiği mikroşerit Sanki-TEM modu	4
Şekil 2. 3. Tipik dalga kılavuzları	5
Şekil 2. 4. Dalga kılavuzu kesme frekansı	5
Şekil 2. 5. Kayıpsız bir malzeme ile doldurulmuş, mükemmel iletken duvarlara sahip dikdörtgen bir dalga kılavuzunun geometrisi	6
Şekil 2. 6. Dikdörtgen dalga kılavuzu TE ₁₀ modu alan dağılımı; a) Elektrik alan çizgileri (önden görünüm); b) Manyetik alanı (üstten görünüm); c) Elektrik alan (üstten görünüm)	7
Şekil 2. 7. Düzlemsel ve düzlemsel olmayan yapılar arasındaki performans boşluğu, YGDK gelişimine yol açtı	8
Şekil 2. 8. (TE ₁₀) modu YGDK iletim hattı; a) Kesit; b) üstten görünüm	9
Şekil 2. 9. YGDK 'nin üstten görünümü	9
Şekil 2. 10. Mikroşerit portu kullanarak YGDK uyarımı; a) (TE ₁₀) -YGDK; b) Sanki-TEM mikroşerit hattı; c) YGDK iletim hattında (TE ₁₀ modu uyarımı Sanki-TEM modunu kullanarak elektrik alan dağılımı)	11
Şekil 2. 11. Konik şekilli YDK- Mikroşerit geçişinin konfigürasyonu	12
Şekil 2. 12. Mikroşerit ve YGDK arasındaki yeni konik geçişin yapısal parametreleri (Kordiboroujeni ve Bornemann, 2014)	12
Şekil 2. 13. Basit ÇTY yuvaları	13
Şekil 2. 14. ÇTY hücre şeması; “Turuncu bakır, beyaz alt tabakanın yalıtım malzemesini gösterir”	13
Şekil 3. 1. Tezin genel tasarım prosesi	14
Şekil 3. 2. Filtre tasarım prosesi	15
Şekil 4. 1. Üçüncü dereceden YGDK-ÇTY bant geçiren filtre; a) Üst düzlem; b) Alt düzlem; “Sarı, bakır ve gri, alt tabakanın yalıtım malzemesini gösterir”	17

Şekil 4. 2. Üçüncü dereceden YGDK-ÇTY C-Bandı bant geçiren filtre; a) S-parametreleri; b) Grup gecikmesi	18
Şekil 4. 3. Üçüncü dereceden YGDK -ÇTY C-Bandı BGS; a) Üst düzlem; b) Alt düzlem; “Sarı bakır ve gri alt tabaka yalıtım malzemesini gösterir”	20
Şekil 4. 4. Üçüncü dereceden YGDK-ÇTY C-Bandı BGS; a) S-parametreleri; b) Grup gecikmesi.....	21
Şekil 5. 1. Fabrikasyon X-Band filtresi; a) Üst düzlem; b) Alt düzlem	22
Şekil 5. 2. SMA konektör ve karşılık gelen boyutları.....	23
Şekil 5. 3. SMA konektörünün hatalı kurulumu; a) üst yüz; b) alt yüz.....	24
Şekil 5. 4. Ölçülen S_{21} yanıtı.....	25
Şekil 5. 5. Ölçüm sonuçlarına karşı ideal benzetim sonuçları	26
Şekil 5. 6. Simülasyona SMA'nın dahil edilmesi; a) üst görünüm; b) alttan görünüm; c) yandan görünüm.....	27
Şekil 5. 7. SMA hatalı kurulumu dikkate alınarak simüle edilmiş S-parametreleri, ölçülen S-parametrelerine karşı.....	27
Şekil 5. 8. SMA yanlış kurulum nedeniyle alan dağılımı; a) E-alan üstten görünüşü; b) E-alanın alttan görünüşü; c) E-alan yandan görünüşü; d) H-alanı üstten görünüşü; e) H alanı alttan görünümü; f) H alanı yandan görünüşü;.....	30
Şekil 5. 9. SMA konektör kurulumunu düzeltme	31
Şekil 5. 10. SMA kurulumunun düzeltilmesinden sonra ölçülen sonuçlar; a) ölçülen S_{21} yanıtı; b) ölçülen S_{11}	32
Şekil 5. 11. Ölçülen S-parametrelerine karşı ideal Benzetim	32
Şekil 5. 12. Simülatöre SMA konektörleri dahil etmek; a) üstten görünüm; b) üst yüzün eğik görünümü; c) alt yüzün eğik görünümü	34
Şekil 5. 13. Benzetim SMA-dahil, Ölçülen sonuçlar karşı	34
Şekil 5. 14. Yüksek performanslı konektörler; a) yüksek performanslı SMA konektörü 26,5 GHz'e kadar hizmet verir; b) 2,92 mm veya k- konektörü 50 GHz'e kadar hizmet eder.....	35

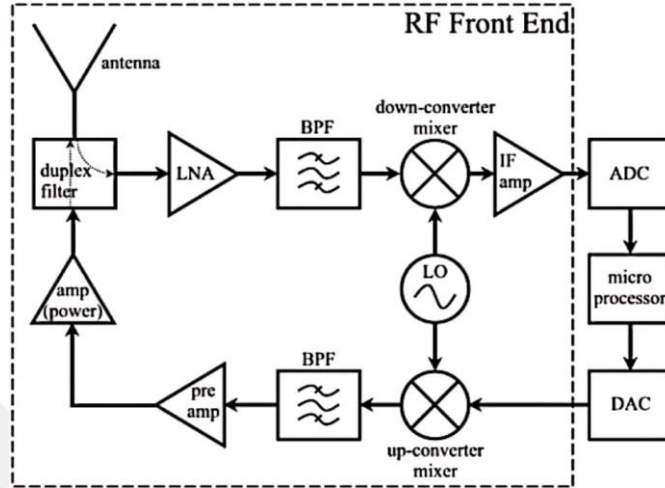
ÇİZELGELER DİZİNİ

Çizelge 2. 1. Dalga kılavuzları, düzlemsel iletim hatları ve YGDK arasındaki özelliklerin karşılaştırılması	8
Çizelge 4. 1. X-Bant filtre boyutları	17
Çizelge 4. 2. C-Bandı filtre boyutları	20



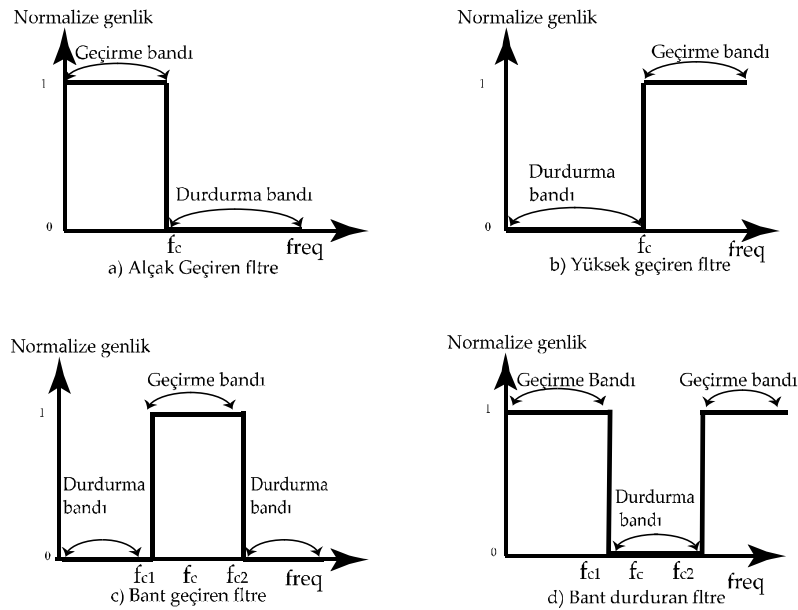
1. GİRİŞ

Filtreler pasif bileşenlerdir ve herhangi bir iletişim sisteminin temel bileşeni olarak kabul edilirler. Filtreler, kalan istenmeyen frekansı (durdurma bandı) bloke ederken istenen sinyalin geçmesini (geçiş bandı) sağlayan iki bağlantı noktalı bir frekans seçici ağılardır. Radar ve iletişim sistemlerinde geniş bir uygulama alanına sahiptirler. Şekil 1.1'de gösterildiği gibi yaygın olarak alıcı-verici devrelerinde kullanılır.



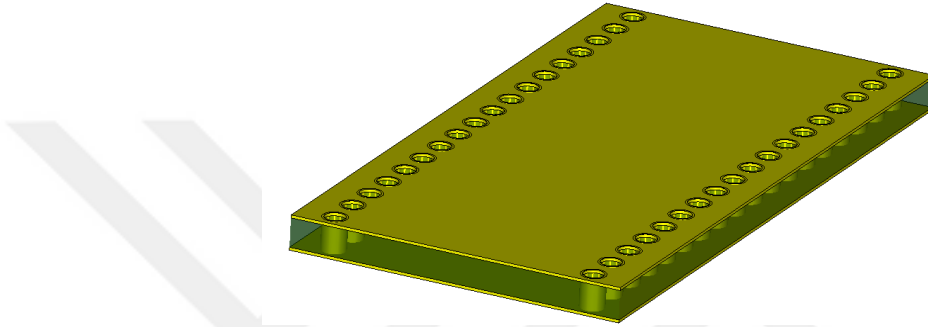
Şekil 1. 1. Basitleştirilmiş RF ön uç alıcı-verici blok şeması

Filtreler, Şekil 1.2'de gösterildiği gibi, frekans seçimi açısından alçak geçiren, yüksek geçiren, bant geçiren ve bant kesen olmak üzere dört kategoride sınıflandırılır. Ayrıca, yanıtı (Chebyshev, Maksimal düz, Eliptik, vb.) Veya teknolojiye (Topaklı, Dalga Kılavuzu, Mikroşerit, vb.) Veya frekans bandına (Dar Bant veya Geniş Bant) göre sınıflandırılabilir.



Şekil 1. 2. Frekans seçiciliğine göre filtre türleri

Geleneksel teknoloji, gerekli tüm özellikleri bir anda karşılayamaz. Geleneksel teknolojiler arasında bir performans açığı var. Geleneksel teknolojiler arasındaki boşluğu doldurmak için gelen çözüm ne olacak? Son zamanlarda, Şekil 1.3'te gösterilen Yama giydirilmiş dalga kılavuzu (YGDK) adı verilen bir düzlemsel yapısına dalga kılavuzu yapılarını sentezleyen tümleşik bir kılavuz yapı. İlk olarak (Hirokawa ve Ando, 1998) tarafından sunuldu ve (Deslandes ve Wu, 2001) tarafından tanıtıldı. YGDK Düzlemsel ve düzlemsel olmayan dalga kılavuzu teknolojileri arasındaki boşluğu doldurabilecek en iyi uzlaşmayı sağlar ve burada hem RWG hem de mikroşerit hatlardan uyarlanmış avantajlardan yararlanır. Ayrıca, düşük kayıp profilli dielektrik altlıklarda PCB'nin imalat prosesleri yoluyla imalat kolaylığı ile.



Şekil 1. 3. Yama giydirilmiş dalga kılavuzunun (YGDK) genel yapısı.

Bu tezde, iki YGDK filtresi tasarlamaya odaklanıyoruz, C ve X bandı üzerinde çalışıyoruz. Biri X-bandı uygulaması için diğeri C-bandı boşluk iletişimi için önerilen iki filtre.

İlk tasarım, YGDK için literatürde bulunan ampirik formüller ve tasarım kriterleri uygulanarak geliştirilmiştir. Bir başka deyişle ilk tasarımın tam dalga elektromanyetik benzetimi ile işe başlanır. YGDK modeli, iki aşamada uygulanan doğrudan EM optimizasyonunu kullanarak optimize edilir: İstenen düşük kesim frekansını elde etmek için YGDK genişliğini optimize edilir ve ardından yansımaları en aza indirmek amacı ile baskın mod bant genişliğini genişletmeye çalışılır.

Takip eden adım hızlı bir şekilde benzetimi tamamlamak ve istenen bant boşluğu etkisini sağlayan hücrelerin başlangıç boyutunu elde etmektir. Bunun için bir mikroşerit iletim hattı altında hücreleri yükleyerek Çentikli Taban Yapıya (ÇTY) uygun hücrelerinin etkilerini araştırılır. Bu süreçlerin tamamlanmasından sonra elde edilen hücreler, bant geçiren filtre oluşturmak için YGDK yapısı ile kullanılacak başlangıç hücreleri olarak kullanılabilir hale gelir.

Son zamanlarda, YGDK ve ÇTY mimarisi kullanılarak gerçekleştirilen önemli sayıda bant geçiren filtre ortaya çıktı, bu filtrelerin bazı örnekleri (L. H. Weng, 2008; Y. Huang, 2013) 'de belirtilmiştir. Giriş daha detaylı şekilde Ek-1 (sayfa 1-5)'te sunulmuştur.

2. KAYNAK TARAMASI

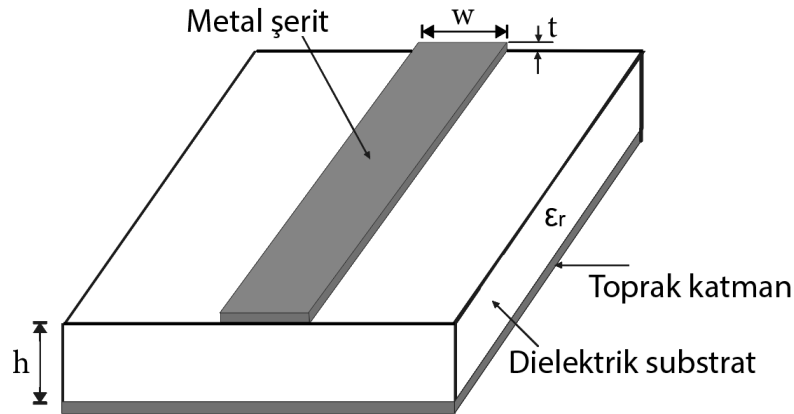
Bu Bölüm aracılığıyla, arızalı zemin yapısı (ÇTY) teorisine ek olarak, bu tez’de kullanılan dalga kılavuz yapıları veya iletim hattı tipleri ile ilgili temel iletim hatları teorisinin bir arka planı sunulacaktır.

2.1. İletim Hatları

İletim hattı devresi, elektrik sinyallerini bir noktadan diğerine taşıyabilen ve yönlendirebilen bir yapıdır. Yayılan EM dalgaları enerjiyi hem elektrik hem de manyetik alanlarda depolarken, iletim hatları elektrik ve manyetik enerjiyi dağıtılmış kapasitans ve endüktans şeklinde depolar (Steer, 2016). Son zamanlarda, YGDK üzerine yoğun araştırmalar, mikrodalga ve milimetre dalga uygulamalarıyla ilişkili devre tasarımı geliştirme seviyelerinin ufkunu artıran ve düzlemsel olmayan devre kavramlarını düzlemsel yapılara uygulamayı sağlayan yeni bir iletim hattı türü olarak ortaya çıkmasına yol açtı. Ayrıca YGDK, dalga kılavuzu performansı ile karşılaştırılabilir performansa sahip olup diğer düzlemsel yapılarla entegre edilmesi kolay düzlemsel bir yapıdır.

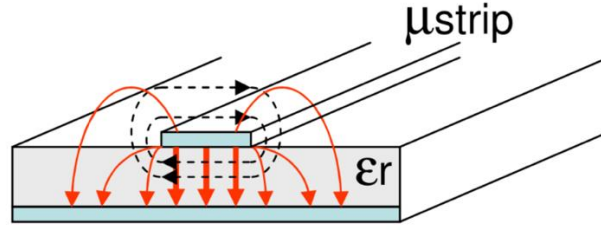
2.1.1. Mikroşerit Hatları (HONG, 2011)

Mikroşerit yapı, ince dielektrik altlığın üst tarafında göreceli geçirgenliğe (ϵ_r) ve altlık kalınlığı “h”, genişliği “W” ve kaplama kalınlığına “t” sahip iletkenlerin oluşturduğu en popüler düzlemsel tip iletim hattıdır, taban ise Şekil 2.1’de gösterildiği gibi iletken tabaka ile kaplıdır.



Şekil 2. 1. Mikroşerit çizgi genel yapısı

Mikroşerit, Şekil 2.2’deki gibi Sanki-TEM modunu destekler, iletim şeridi ile toprak arasında açılan homojen olmayan yapısından dolayı hava (saçak alanı) ve dielektrik olmak üzere iki farklı ortamda bulunur.

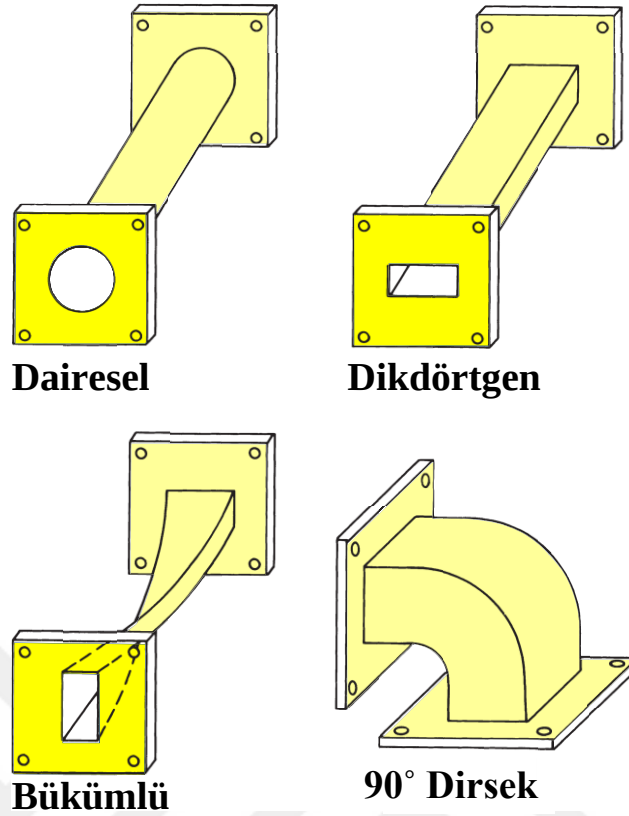


Şekil 2. 2. Kesikli çizgilerin manyetik alan çizgilerini ve düz çizgilerin elektrik alan çizgilerini temsil ettiği mikroşerit Sanki-TEM modu

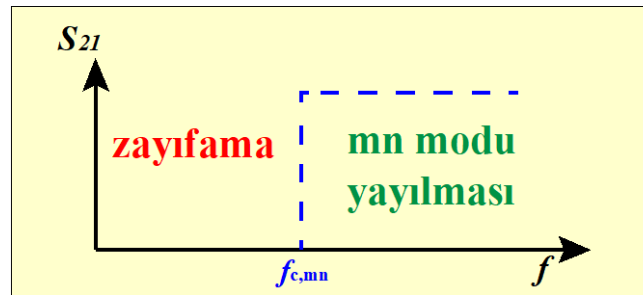
Mikroşerit hat bileşenleri, yekpare boyuta ek olarak tasarım, imalat, entegrasyon kolaylığı avantajına sahiptir. TEM modu için bir kesme frekansı yoktur. Mikroşerit hatlar doğru akımdan (DC) yüksek frekans bölgesine kadar işaret iletebilir. Söz konusu özellikler RF ve mikrodalga devre tasarımında yaygın olarak kullanılan teknolojinin yapılmasına yol açmıştır. Diğer yandan, mikroşerit yapılar yüksek kayıplar ve düşük güç destekleyebilme özelliklerinden mustarıdır. Bu arada mikroşerit yapılar sınırsız yapıları nedeniyle çapraz konuşma ve kasıtsız radyasyona karşı hassastır. Daha fazla detay ve mikroşerit denklemleri Ek-1 (sayfa 6-8)'de sunulmuştur.

2.1.2. Dikdörtgen Dalga Kılavuzları

Metal dalga kılavuzları tipik olarak Şekil 2.3'te gösterildiği gibi bir yalıtım ortamına yerleştirilmiş kapalı bir iletken yapıdır. Dalga kılavuzlarının çoğu hava dolu tiptir, diğer dielektrik dolu dalga kılavuzlarından daha iyi performansa sahiptir, daha yüksek dielektrik sabitinin diğer dielektrikleri boyut küçültme amaçları için kullanılır. Dalga Kılavuzunda kullanılan dielektriğin nispi dielektrik sabiti arttığında Dalga Kılavuzunun boyutunun küçülmesidir. Tek tümleşik bir iletken bir yapıdan oluştukları için TEM modunu destekleyemezler. Ancak dalga kılavuzu enine elektrik (TE) veya enine manyetik (TM) modları ile bunların hibrit modları adı verilen bir kombinasyonu destekler. İşaretler ancak baskın TE veya TM modu kesme frekansının üzerinde ise taşınabilir (Pozar, 2012). Burada, belirli bir çalışma frekansında f , Şekil 2.4'te gösterildiği gibi sadece $f > f_c$ olan modlar yayılabilir. F_c kesme frekansı, altında zayıflamanın gerçekleştiği ve üzerinde yayılımın gerçekleştiği frekanstır (Yüksek Geçiş).



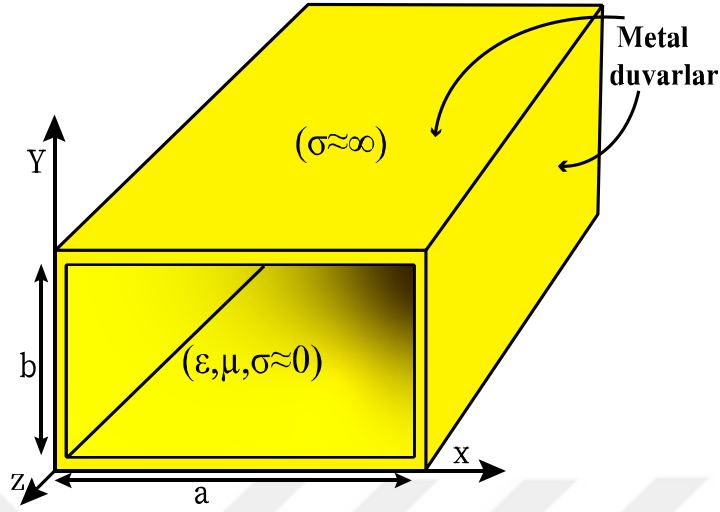
Şekil 2. 3. Tipik dalga kılavuzları



Şekil 2. 4. Dalga kılavuzu kesme frekansı

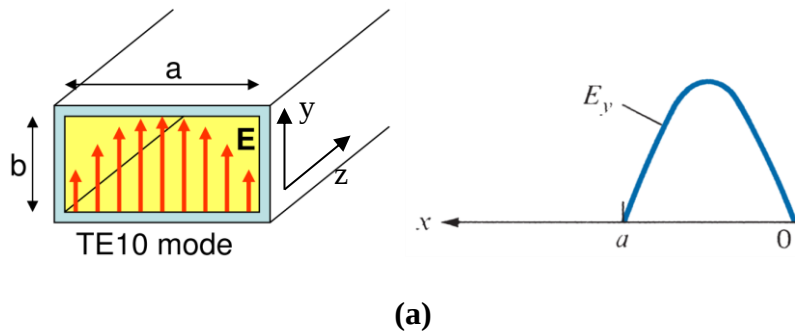
Dalga kılavuzları, iletim hatlarından daha düşük sinyal zayıflaması ve daha büyük bant genişliği sağladığı için yüksek frekanslarda tercih edilir. Ayrıca, yayılım dalgaları alanları, küçük kesitli iletim hatlarına nazaran daha geniş bir kesit alanı üzerinden daha düzgün yayıldığından, yüksek güç seviyelerini destekleme kabiliyeti için gerekli olan yüksek Q faktörü elde etmek kolaydır. Bununla birlikte, dalga kılavuzları büyük boyut ve yüksek üretim maliyetleri nedeniyle pratik değildir, ayrıca düzlemsel devreleri ile entegrasyon problemlerinden mustarıdır.

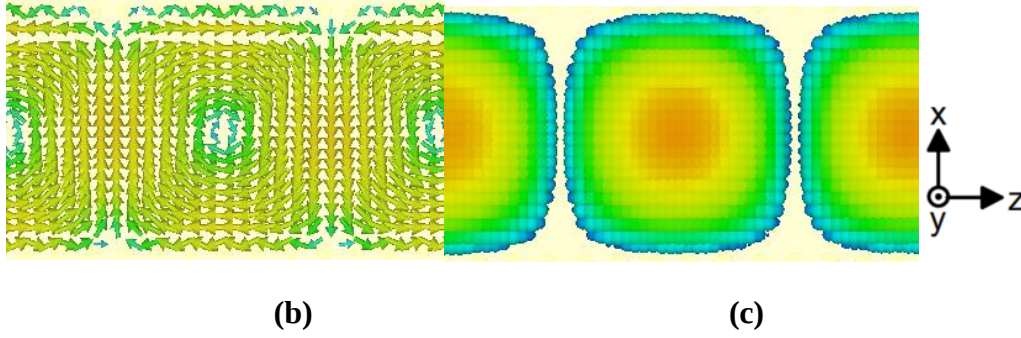
En çok kullanılan dalga kılavuzları dikdörtgen kesitli dalga kılavuzlarıdır. Tipik yapı Şekil 2.5'te gösterilmiştir ve burada a ve b iç boyutlardır.



Şekil 2. 5. Kayıpsız bir malzeme ile doldurulmuş, mükemmel iletken duvarlara sahip dikdörtgen bir dalga kılavuzunun geometrisi

TE_{10} modunun en düşük kesme frekansına sahip baskın mod olduğu ($a > b$) standart bir uygulamadır. TE_{10} modu şu özelliklere sahiptir: (1) elektrik alanı yayılma yönüne dik düzlemlerle sınırlıdır, manyetik alan hem enine hem de boyuna bileşenlere sahiptir ve (2) elektrik alanı düz kılavuz yüzeylerinden birinden yönlendirilir Şekil 2.6.a'da gösterildiği gibi zıt yüzeyin karşısında. Şekil 2.6.b, 2.6.c, dikdörtgen dalga kılavuzunda TE_{10} modunun elektrik ve manyetik alanlarının benzetimi yapılmış dağılımının üstten görünüşünü göstermektedir. Dikdörtgen Dalga Kılavuzları ile ilgili fazla detay ve denklemleri Ek-1 (sayfa 9-12)'de sunulmuştur.

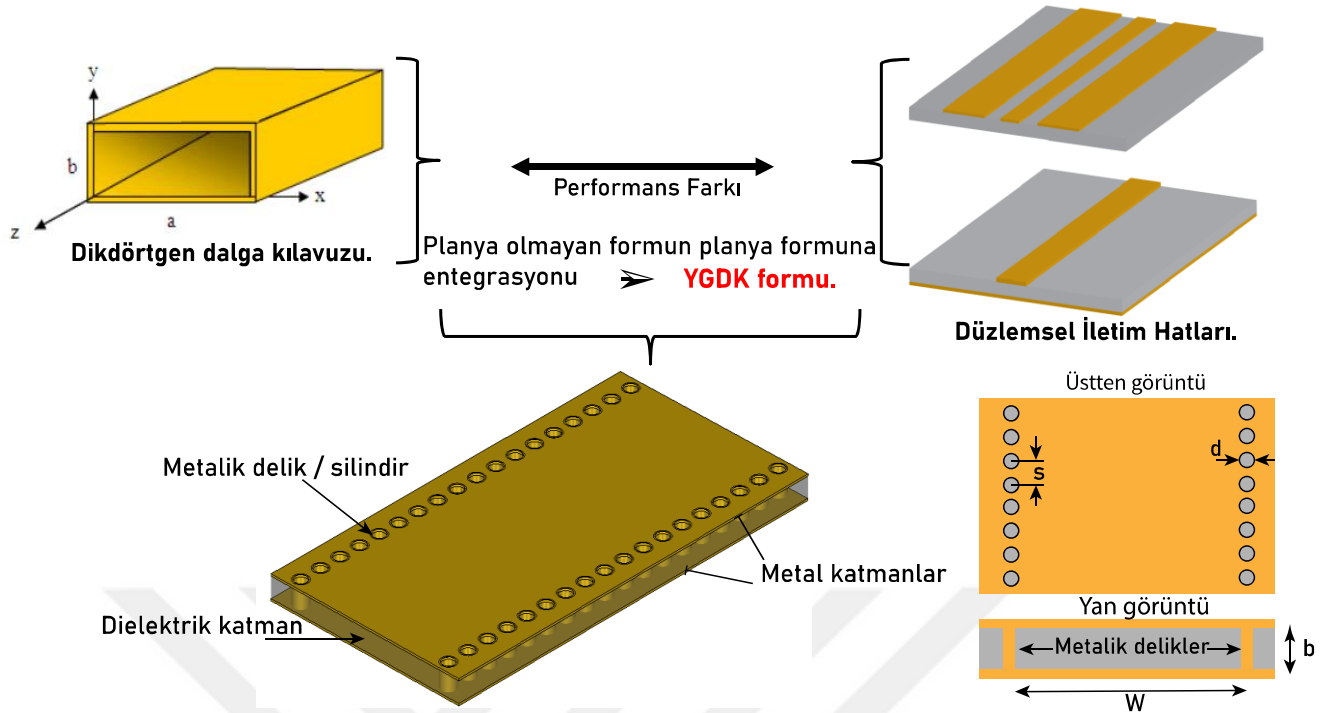




Şekil 2. 6. Dikdörtgen dalga kılavuzu TE_{10} modu alan dağılımı; **a)** Elektrik alan çizgileri (önden görünüm); **b)** Manyetik alanı (üstten görünüm); **c)** Elektrik alan (üstten görünüm)

2.1.3. Yama Giydirilmiş Dalga Kılavuzu (YGDK)

Yama Giydirilmiş Dalga Kılavuzu (YGDK) baskı devre altlıklarını kullanarak ortaya çıkmış bir tür dalga kılavuzu yapılarıdır. Bu yeni iletim hattı son on yılda popüler hale geldi, YGDK prensibi dielektrik dolgulu dikdörtgen dalga kılavuzundan kaynaklandı. YGDK'nin geliştirilmesinin arkasındaki gereklilik, dalga kılavuzlarının yüksek güç kapasitesi ve yüksek Q faktörü sunduğu geleneksel teknolojiler arasındaki boşluktur, ancak hantal ve pahalıdır, ayrıca düzlemsel devreleri ile entegrasyon zorlukları vardır. Diğer taraftan, düzlemsel iletim hatları, diğer düzlemsel devreleriyle entegrasyon kolaylığı sağlayan düşük profilli (düşük profilli küçük yükseklik ve genişlik anlamına gelir) yapılarıdır. Ancak yüksek kayıplar ve radyasyon sızıntısı olan yüksek performanslı gereksinimlerden yoksundurlar. YGDK, metalik dikdörtgen dalga kılavuzunun düzlemsel bir mikroşerit alt tabakaya bir tür entegrasyonudur, tek katmanlı yapılar için baskılı devre kartı (BDK) içeren standart düzlemsel imalat işleme teknikleri kullanılarak oluşturulabilir veya düşük sıcaklıkta pişirilmiş seramik (DSPS) ve (liquid crystal polymer) sıvı kristal polimer (LCP) gibi çok katmanlı yapılar için diğer işleme teknikleri kullanılabilir. Şekil 2.7 dikkatle incelendiğinde YGDK'nin dikdörtgen bir dalga kılavuzunun bir mikroşerit alt tabakaya doğrudan bir entegrasyonu olduğunu söyleyebiliriz. Bu, mikroşerit teknolojisi ve Hava dolu dikdörtgen dalga kılavuzu teknolojisi arasında en iyi uzlaşmayı sağlar, YGDK, dielektrik dolum yapısının yanı sıra hacim azaltımı nedeniyle Hava dolu dalga kılavuzunun Q faktörünü azaltır, ancak düzlemsel devre dahil tüm devre, geçiş ve dalga kılavuzu hepsi bir plaka üzerindedir (Deslandes ve Wu, 2001). Bu son zamanlarda akademisyenleri YGDK geliştirmeye motive etti (Ke, Deslandes ve Cassivi, 2003). İyi imal edilmiş bir YGDK, Terahertz Uygulamalarına (Wang, Hao ve Kui-Kui, 2016) hizmet vermesini sağlayan 150 GHz'e kadar çalışabilir, ayrıca üretim yeteneğine bağlı olarak daha yüksek frekanslara kadar çalışabilir.



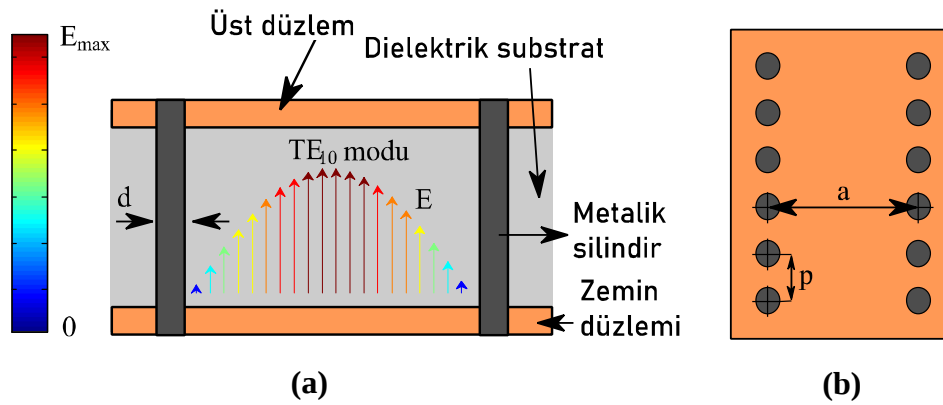
Şekil 2. 7. Düzlemsel ve düzlemsel olmayan yapılar arasındaki performans boşluğu, YGDK gelişimine yol açtı

Bu yeni mimari, Tablo 2.1'de gösterildiği gibi hem mikroşerit hem de dikdörtgen dalga kılavuzundan özellikleri birleştiren özelliklere sahiptir.

Çizelge 2. 1. Dalga kılavuzları, düzlemsel iletim hatları ve YGDK arasındaki özelliklerin karşılaştırılması

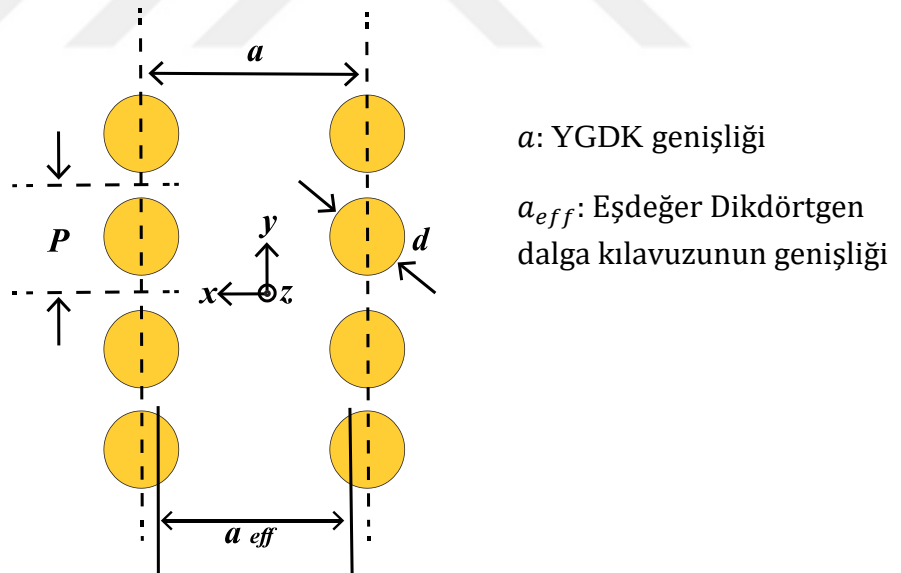
	Maliyet	Kayıp	Güç kullanma kapasitesi	Yekpare boyut	Öz-ekranlanmış	Öz-ambalaj
Dalga kılavuzları	☹️	😊	😊	☹️	😊	😊
Düzlemsel İletim hatları	😊	☹️	☹️	😊	☹️	☹️
YGDK	😊	😊	😊	😊	😊	😊

Prencip olarak, Yama giydirilmiş dalga kılavuzu (YGDK), geleneksel dikdörtgen dalga kılavuzlarının temel prensiplerine benzer prensiplere bazı farklılıklarla uyar. YGDK'nin baskın modu, Şekil 2.8'da gösterildiği gibi (TE_{10}) 'dur.



Şekil 2. 8. (TE₁₀) modu YGDK iletim hattı; **a)** Kesit; **b)** üstten görünüm

YGDK tasarımımda, belirlenen parametreler sırasıyla g silindirin çapı “d”, bitişik yollar arasındaki merkezden merkeze mesafe “p” ile temsil edilmektedir ve YGDK genişliği, Şekil 2.9'da gösterilmiştir. Bu parametreler, sızıntı kayıplarını önlemek ve periyodik yapılarda ortaya çıkan bant aralığı “durdurma bandı” etkilerinden kaçınmak için dikkatle seçilmeli, ayrıca yapının sağlamlığını korumak ve aşırı orantıdan kaçınmak için bunların hepsinden kaçınmak için seçilmelidir. (Deslandes ve Ke, 2006) tarafından belirtilen bu efekt tasarım kurallarına uyulmalıdır;



Şekil 2. 9. YGDK 'nin üstten görünümü

YGDK'nın mekanik tokluğunu (Aşırı delinmesini önlemek için) sağlamak Denklem 2.23 ile verilen koşulun sağlanması gerekir.

$$P/\lambda_c > 0.05 \quad (2.23)$$

Herhangi bir bant boşluğu efektinden kaçınmak için Denklem 2.24 ile verilen koşulun sağlanması gerekir.

$$P/\lambda_c < 0.25 \quad (2.24)$$

Alan sızıntısını önlemek için Denklem 2.25 ile verilen koşulun sağlanması gerekir.

$$P \leq 2d \quad (2.25)$$

YGDK bir dikdörtgen dalga kılavuzuna eşdeğerdir ve sadece YGDK 'nin etkin genişliği kullanılarak dikdörtgen bir dalga kılavuzu olarak analiz edilebilir (Y. Cassivi, 2002). Bu durumda eş değer dikdörtgen dalga kılavuzunun genişliği Denklem 2.26 ile tanımlanmaktadır.

$$a_{eff} = a - \frac{d^2}{0.95 * p} \quad (2.26)$$

Baskın mod kesim frekansı (TE_{10}) ifadesinin Denklem 2.28 ile verilen koşulu sağlanması gerekir.

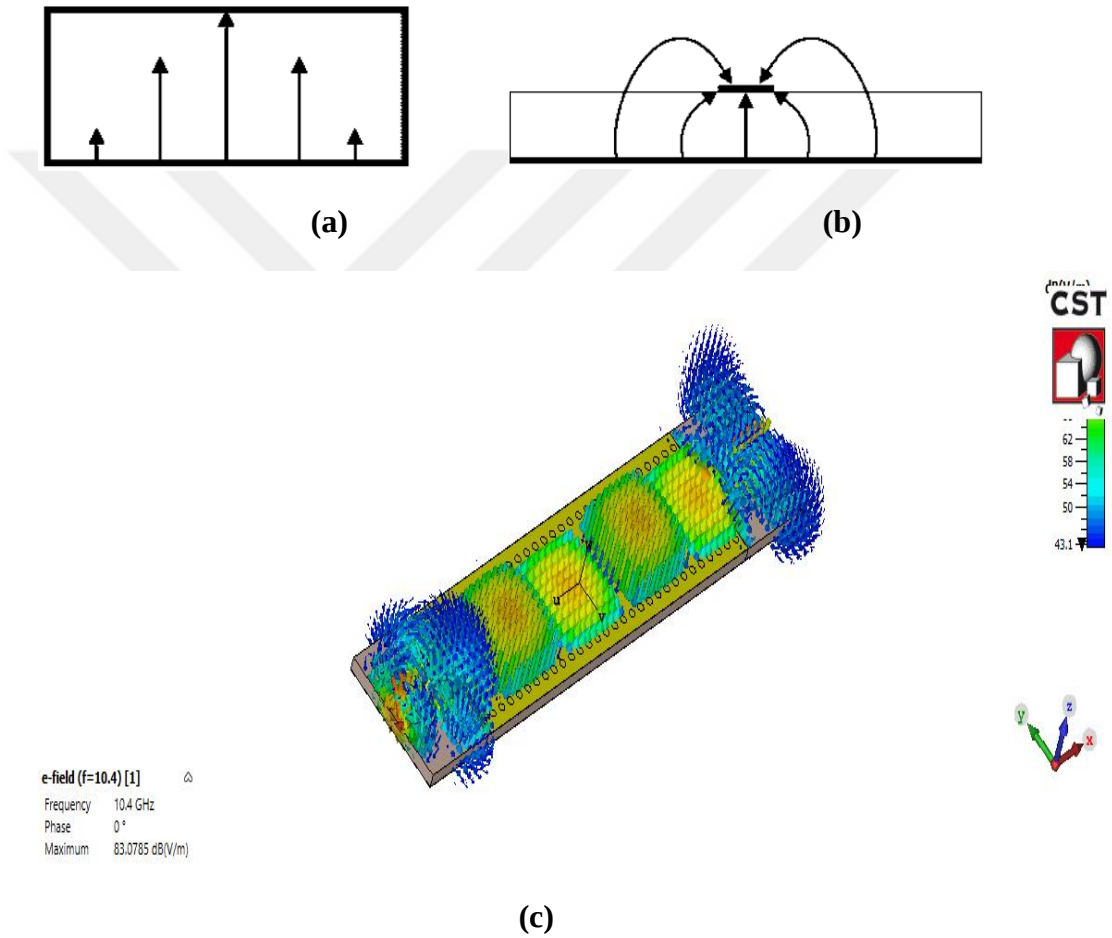
$$F_{c(TE_{10})} = \frac{C_0}{2 \cdot \sqrt{\epsilon_r}} \left(a - \frac{d^2}{0.95 * b} \right)^{-1} \quad (2.28)$$

Bu denklemde, C_0 ışık hızını, d metalik silindirin çapı, p bitişik metalik silindirler arasındaki merkezden merkeze mesafe ve a YGDK'nın genişliği ifade etmektedir. Dikkat edilmelidir ki YGDK kalınlığı sadece Q-faktörünü etkilemektedir. YGDK ile ilgili daha fazla detay bilgi Ek-1 (sayfa 13-18)'de sunulmuştur.

2.2. Mikroşerit – YGDK geçişi

Filtre, kuplör, güç bölücü ve anten gibi pasif YGDK devrelerini veya amplifikatörler, mikserler, osilatörler gibi aktif devreleri mikroşerit veya ortakdüzlem dalga kılavuzu (ODK: İngilizcesi, CPW) gibi diğer düzlemsel devreleri ile bağlamak ve entegre etmek için Düşük yansıma geçişleri gereklidir. Ayrıca, bileşen tasarımının geçişin etkisinden bağımsız hale getirilmesi gerekmektedir.

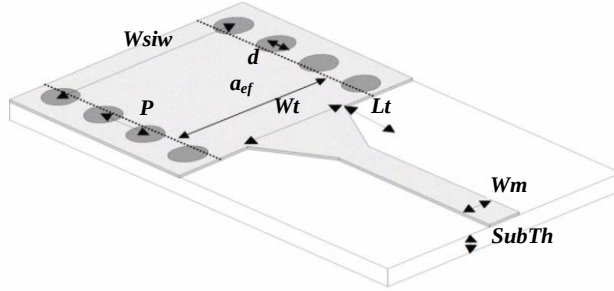
Mikroşerit-YGDK geçişi, Şekil 2.10'de gösterildiği gibi mikroşerit hattı Sanki-TEM modunu kullanarak empedansı uyarlamak ve YGDK temel modunu (TE_{10}) uyarlamak için kullanılır.



Şekil 2. 10. Mikroşerit portu kullanarak YGDK uyarımı; **a)** (TE_{10}) -YGDK; **b)** Sanki-TEM mikroşerit hattı; **c)** YGDK iletim hattında (TE_{10} modu uyarımı Sanki-TEM modunu kullanarak elektrik alan dağılımı)

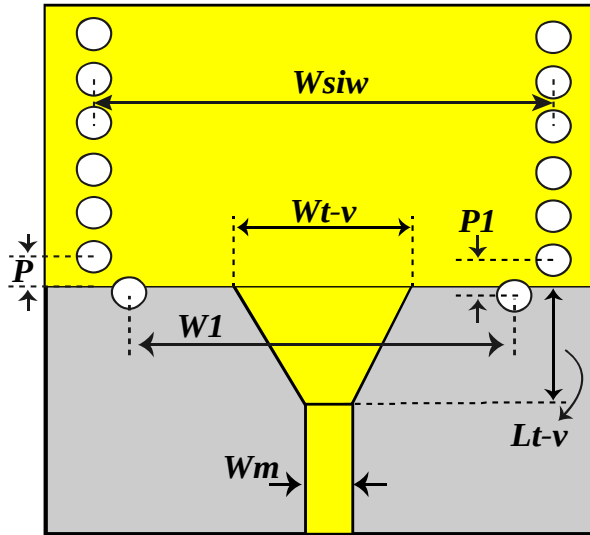
İlk geçiş (Deslandes ve Wu, 2001) tarafından sunulan mikroşerit konikti ve hala Şekil 2.11'de gösterilen tek katmanlı devrelerde en yaygın kullanılan mikroşerit- YGDK geçişi tipiydi. Geniş bant YGDK filtrelerinin tüm bant genişliğini kapsadığı için. Performansı, doğrudan geçiş gibi diğer mikroşerit geçişlere kıyasla daha iyidir ve

tasarımı, Yarı-düzlemsel dalga kılavuzu- YGDK geçişlerinden daha kolaydır, ayrıca iyi optimize edilmiş konik geçiş, minimum ekleme ve geri dönüş kayıpları sağlayabilir. Çoğu zaman, konik geçiş fiziksel boyutları, doğru analitik temsil sağlama zorluğu nedeniyle Elektromanyetik benzetim programlarında bulunan hesaplamalı çözümler kullanılarak hesaplanır (Caleffo, 2016). YGDK (enine elektrik alanı) ve mikroşerit (enine elektromanyetik alan) arasındaki alan modlarındaki fark, doğru analitik çözümler için hesaplama zorluğunu artıran ana faktör olarak kabul edilir.



Şekil 2. 11. Konik şekilli YDK- Mikroşerit geçişinin konfigürasyonu

Mikroşerit Hattı- Şekil 2.12'te gösterilen (Kordiboroujeni ve Bornemann, 2014) tarafından sunulan yama giydirilmiş dalga kılavuzu yeni geniş bant geçişi, bu geçiş, geleneksel geçişten (Deslandes ve Wu, 2001) daha iyi seviyede bir ekleme ve geri dönüş kayıpları sunmaktadır. Mikroşerit – YGDK geçişi ile ilgili daha detay Ek-1 (sayfa 19-21)'de sunulmuştur.

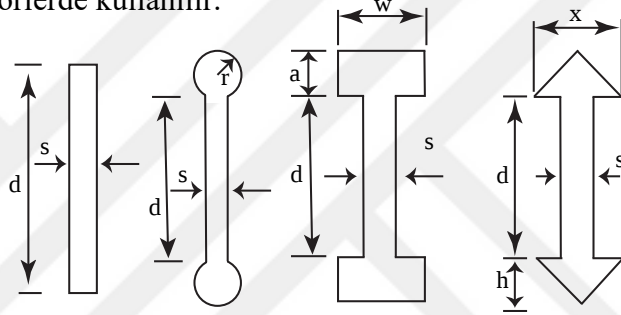


Şekil 2. 12. Mikroşerit ve YGDK arasındaki yeni konik geçişin yapısal parametreleri (Kordiboroujeni ve Bornemann, 2014)

2.3. Çentikli Taban Yapılar (ÇTY)

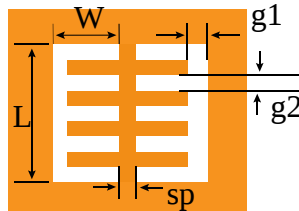
Çentikli Taban Yapılar (ÇTY) ilk olarak Fotonik Bant Aralığı (FBA) yapılarının incelenmesine dayanarak Kore bilim adamı (Jong-Im ve diğerleri, 1999) tarafından önerilmiştir. ÇTY, bazen daha karmaşık şekiller kullanarak, zemin düzleminde basit bir “hata” yapısının aşınmasıyla gerçekleştirilir. Ayrıca, ÇTY daha iyi performans için basamaklı biçimde kullanılabilir. İlk olarak mikroşerit ve yarı-düzlemsel hatlar gibi iletim hatlarının topraklarında kullanılmıştır. Bu yüzden ÇTY olarak adlandırıldı. ÇTY tekniği genellikle bant reddetme özelliği sağlar (Ahn ve ark., 2001).

Basit popüler ÇTY hücrelerinden bazıları, Şekil 2.13'te gösterilen kare kafa yuvası, dairesel kafa yuvası, iki başlı takoz için yuva ve ok ucu yuvasını içerir. ÇTY hücreleri, bant genişliğini arttırmak ve kazancı arttırmak için antenler gibi mikrodalga devrelerinde, ayrıca boyut küçülmesiyle performansı arttırmak için filtrelerde, güç bölücülerinde ve kuplörlerde kullanılır.



Şekil 2.13. Basit ÇTY yuvaları

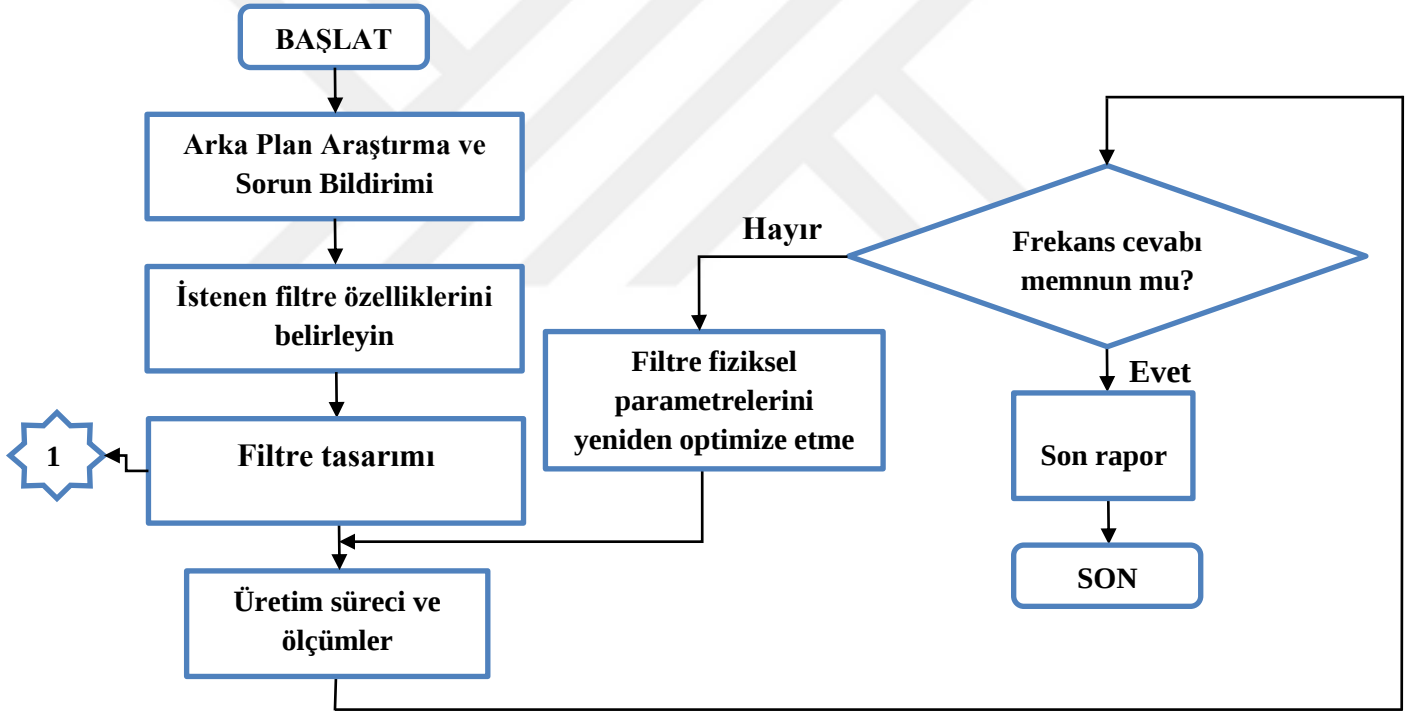
ÇTY hücreleri elektromanyetik bant aralığı sağlar, bu özellik mikrodalga devrelerinde bant duruşlarını ve düşük geçişli filtre yanıtını gerçekleştirmek için kullanılır. (Z. He, 2013) tarafından önerilen yeni bir ÇTY birim hücresi, geleneksel ÇTY hücrelerine göre daha iyi performans sunar, daha fazla kabul edilen YGDK yapılarının üst düzlemine dahil edilebildiği geleneksel hücrelerden ayırt edilmesi avantajına sahiptir. Zemin düzleminde dağlamadan daha iyi performans vermek için YGDK ile uyumludur. Şekil 2.14 hücrenin temel yapısını göstermektedir. Daha teorik detaylar, ÇTY'nin modellenmesi ve uygulamaları Ek-1 (sayfa 21-26)'te sunulmuştur.



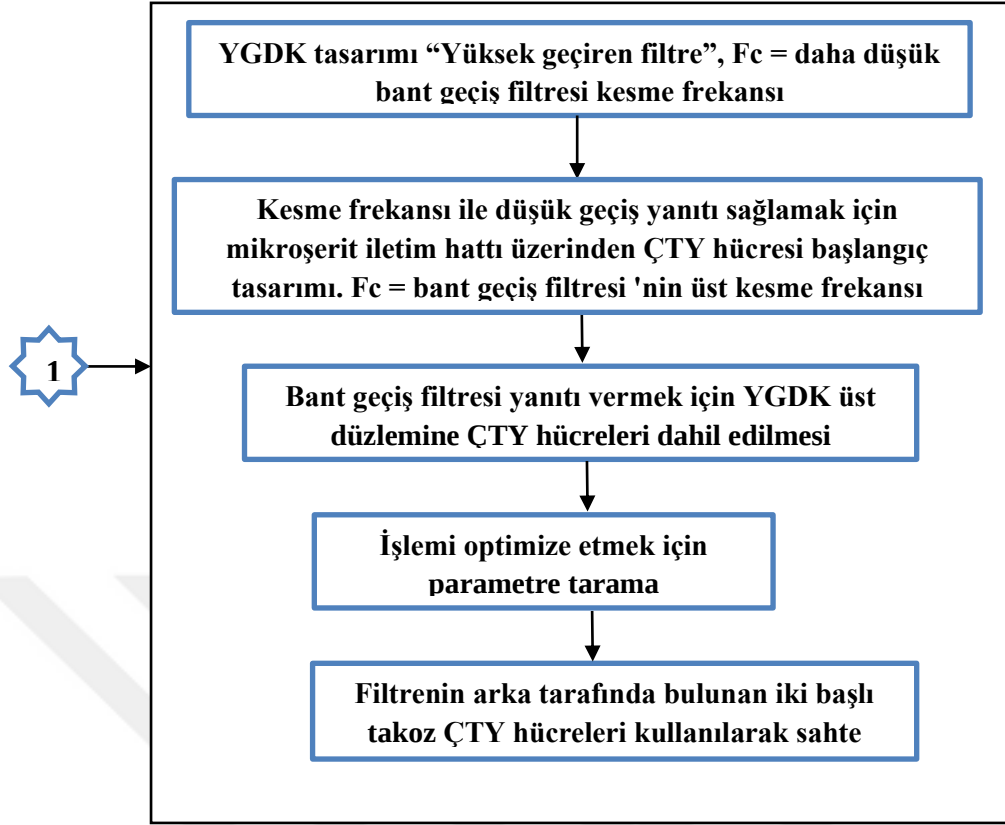
Şekil 2.14. ÇTY hücre şeması; “Turuncu bakırı, beyaz alt tabakanın yalıtım malzemesini gösterir”

3. YÖNTEM

Bu bölümde kısaca filtrelerin genel tasarım stratejisini sunuyoruz. Aşağıda sunulan akış şeması, tasarım sürecinde izlenen genel tasarım yöntemini tanımlar. Bu tez için iki yekpare YGDK -ÇTY bant geçiren filtre tasarlandı, birincisi 4 GHz bant genişliği ile 10 GHz'lik bir merkez frekansa ve (FBW = 40%) kesirli bant genişliğine sahipken, ikinci tasarımın merkezi frekansı Uzay Uygulaması için tasarlanmış 500 MHz bant genişliğine (FBW = 8.3%) sahip 6.175 GHz. Tasarımda kullanılan baskılı devre kartı (PCB), Diclad 880 dokuma fiberglas / PTFE kompozit malzemedir ve dielektrik sabiti 2.2 olup, 10 GHz'de 0.0009 kayıp faktörü ve 0.508 mm kalınlığındadır. Genel tasarım süreci CST- tam dalga simülatörü, zaman alanı çözücüsü kullanılarak ve frekans alanı çözücüsü tarafından da doğrulanmıştır. Ardından, tasarlanan filtrelerin performansını analiz etmek için üretim ve ölçüm yapılır. Şekil 3.1. tezin genel tasarım prosesini gösterirken, Şekil 3.2. filtre tasarım prosesini gösterir. Yöntem daha ayrıntılı Ek-1(sayfa 27-29)'de sunulmuştur.



Şekil 3. 1. Tezin genel tasarım prosesi

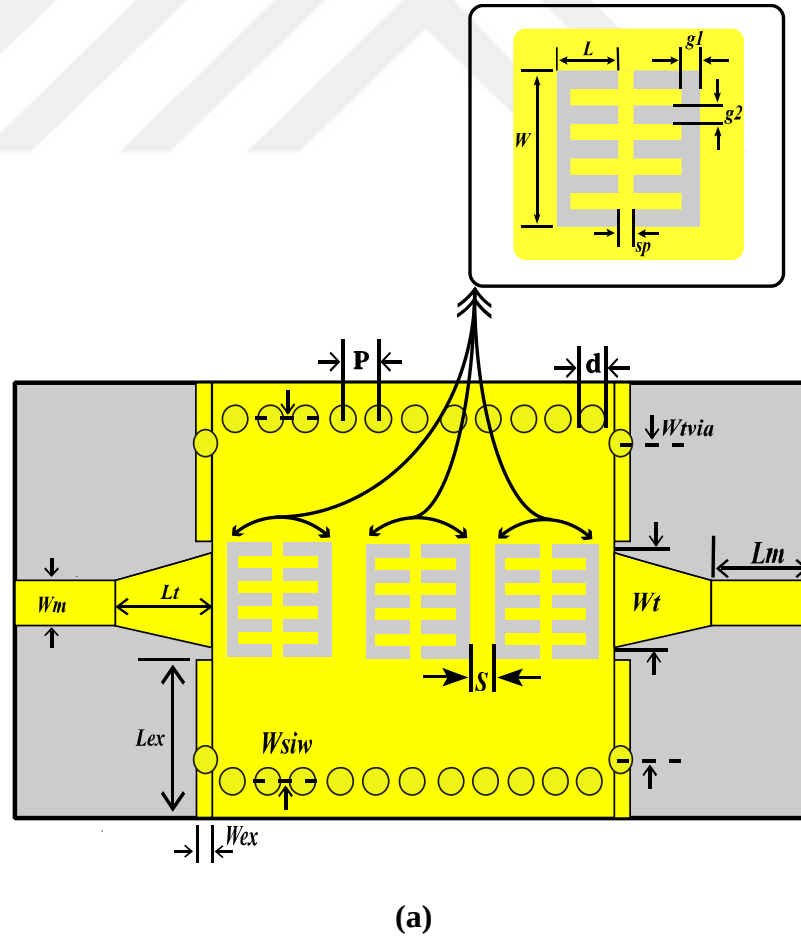


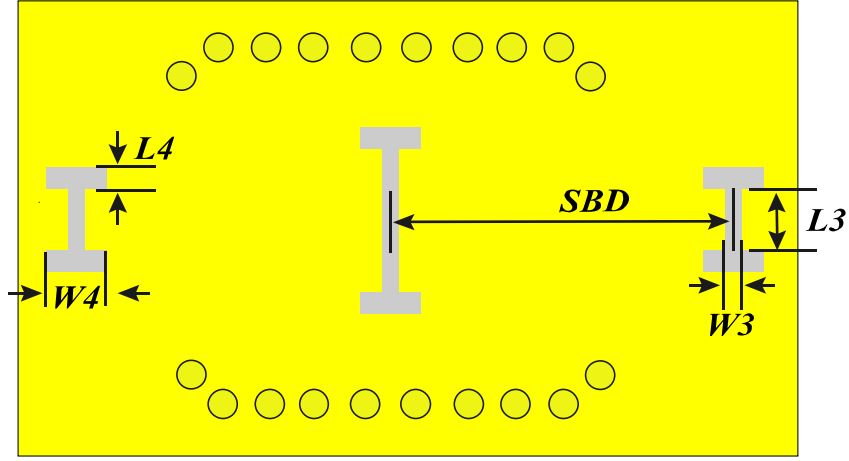
Şekil 3. 2. Filtre tasarım akış diyagramı

4. TASARIM VE BENZETTİM SONUÇLARI

4.1. X– Bant Filtresi

Geniş bant X-band filtrenin merkez frekansı 10 GHz'dir ve kesirli bant genişliği %40'tır, tasarım $\epsilon_r = 2.2$ dielektrik sabitine sahip Diclad 880 altlığı üzerine inşa edilmiştir, dağılma faktörü $\tan \delta = 0.0009$, ve kalınlığı 0.508 mm'dir. Daha önce de belirtildiği gibi, YGDK yüksek geçirgen bir filtre olarak çalışır. ÇTY hücreleri bant aralığı sağlayabilir ve iletim hattı ile birleştirilirse düşük geçişli bir filtre görevi görebilir. Burada her iki kavram da bir bant geçiren filtre tasarlamak için birleştirilir. Ekstra modifikasyonla (Kordiboroujeni ve Bornemann, 2014) konik uzatma, mikroşerit hattının yarı TEM modunu kullanarak YGDK'daki TE10 modunu uyarmak için kullanılır. (Z. He, 2013) tarafından sokulan 3 basamaklı ÇTY hücresi, YGDK'nın üst yüzeyi üzerine kazınmıştır ve bant dışı reddini arttırmak ve ilk olarak tanıtılan ÇTY'nin geçiş bandında 3 hücreli daha iyi geri dönüş kaybı için, filtrenin alt yüzeyinde aşındırılır, biri doğrudan orta üst ÇTY'nin altındadır, diğer 2 hücre mikroşerit portlarının altında aşındırılır. Şekiller 4.1.a ve 4.1.b, bu konfigürasyonun üst ve alt düzlemini gösterirken, Şekiller 4.2.a ve 4.2.b, benzetimi yapılmış S-parametrelerini ve filtrenin iki portu arasındaki Grup gecikmesini gösterir. Teorik hesaplama ve yazılım optimizasyonu filtresinin boyutları Çizelge 4.1'de verilmiştir. CST MW stüdyosunun zaman alanı çözücüsü kullanılarak gerçekleştirilen benzetim süreci.



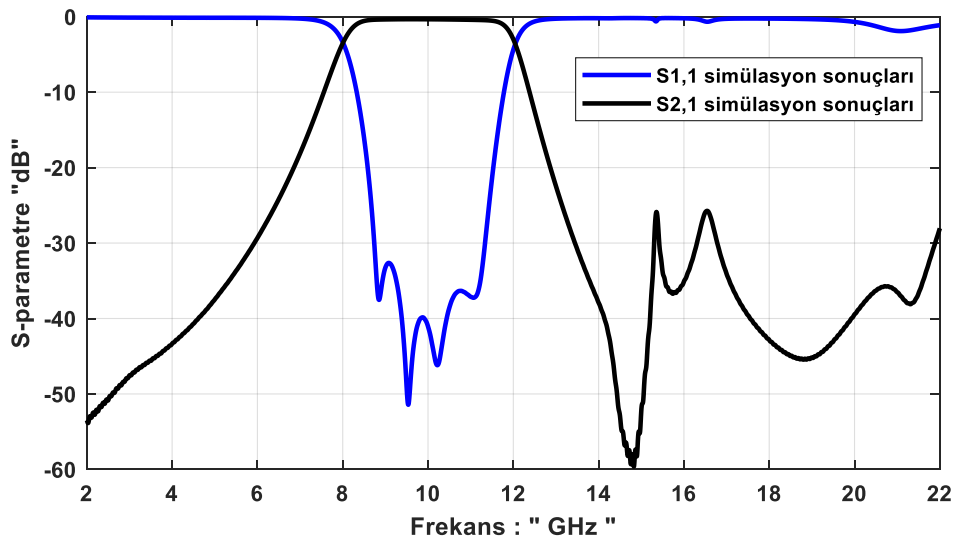


(b)

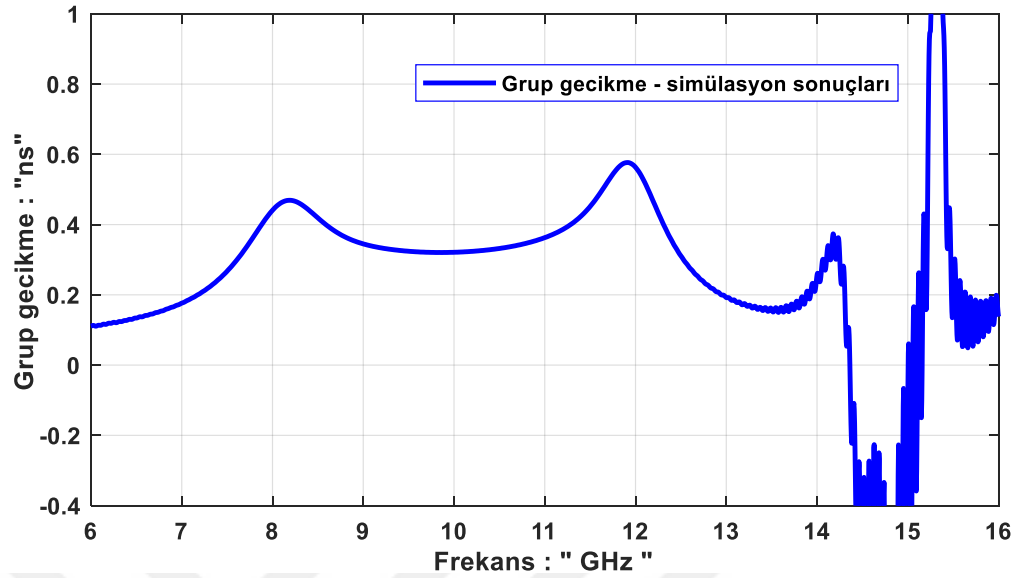
Şekil 4. 1. Üçüncü dereceden YGDK-ÇTY bant geçiren filtre; a) Üst düzlem; b) Alt düzlem; “Sarı, bakır ve gri, alt tabakanın yalıtım malzemesini gösterir”

Çizelge 4. 1. X-Bant filtre boyutları

Değişken	Uzunluk (mm)	Değişken	Uzunluk (mm)	Değişken	Uzunluk (mm)	Değişken	Uzunluk (mm)
Wm	1.919	Wsiw	13.9	SP	0.5	L3	1.7
Lm	3.6	d	1	S	1	W3	0.35
Wt	4.187	P	1.9	L1	4.7	L4	0.5
Lt	3.6	Wtvia	12.4	W1	0.35	W4	0.9
Wex	0.5	P1	1	L2	0.5	SBD	12.5
Lex	6.6	SubTh	0.508	W2	0.9		



(a)



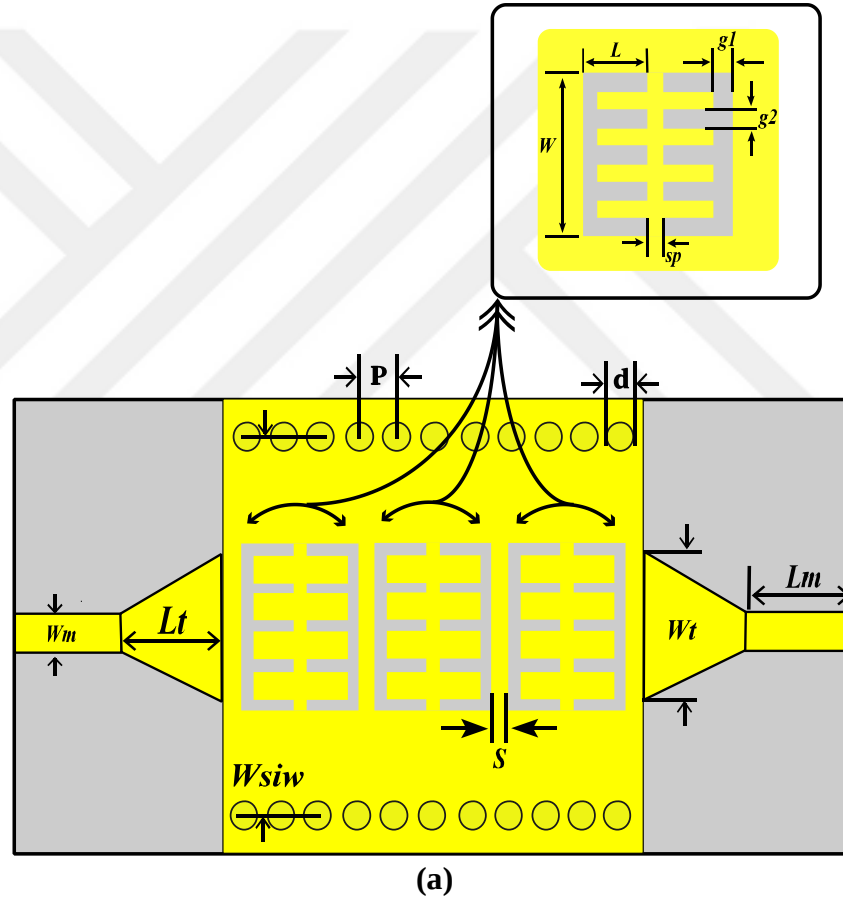
(b)

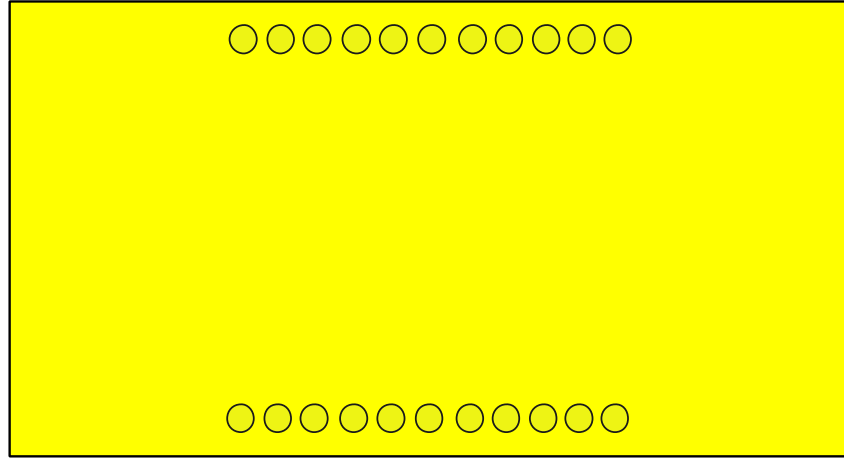
Şekil 4. 2. Üçüncü dereceden YGDK-ÇTY C-Bandı bant geçiren filtre; **a)** S-parametreleri; **b)** Grup gecikmesi

Tablo 4.1'de belirtilen boyutlar, Şekil 4.12'deki optimize edilmiş sonuç için (benzetim ile) nihai değerlerdir. a, 10 GHz merkez frekansında yaklaşık 0.35 dB yerleştirme kaybı (S_{21}) ve dönüş kaybını (S_{11}) geçiş bandında -32 dB'den daha iyi görebiliriz, üst bant reddi ile ilgili olarak geçiş bandında neredeyse düz olan iletim katsayısı 13 GHz'den 22 GHz'den daha fazla 25 dB geniş bant dışı reddine sahipse, sahte yanıt (15.3 ve 16.5) GHz'de 25 dB'den önemli ölçüde düşüktür ve 14.7'de -58dB reddi sıfırlanır. GHz, filtre portları arasındaki grup gecikmesi için Şekil 4.12'de gösterilmiştir. B, 10 GHz merkez frekansında yaklaşık 0.32 ns gecikme ile geçiş bandında yaklaşık olarak düz olduğunu ve varyasyonun oldukça olduğunu fark edebiliriz. Mikrodalga filtre için pürüzsüz. Son olarak, filtrenin boyutu ile ilgili olarak, filtre, toplam boyutu $17.7 * 30 \text{ mm}^2$ olan minyatürleştirme özelliğini korur. Ayrıntılı tasarım ve en iyileştirme aşamaları Ek-1 (sayfa 29-40)'de sunulmuştur.

4.2. C– Bant Filtresi

Amacımız, aşağıdaki özelliklere sahip bir bant geçiren filtrenin basit yekpare düzlemsel tasarımını, 500 MHz bant genişliğine sahip 6.175 GHz'lik bir merkez frekansı ($FBW = \%8,3$) sunmaktır. Önerdiğimiz filtre, bölüm 4.2'de kullanılan aynı hücre yapısı ile aynı alt tabaka ile 3 basamaklı ÇTY hücresi ile yüklenen YGDK yapısını birleştiren Üçüncü dereceden YGDK-ÇTY C-Bant filtresini oluşturur. Ortaya çıkan filtre yapısı, Şekil 4.3.a, b'de gösterilmiştir ve boyutları, Tablo 4.2'de verilmiştir. Filtre, CST MW tam dalga simülatörü kullanılarak modellenmiştir ve benzetim, zaman alanı çözücü kullanılarak gerçekleştirilmiştir. Benzetim sonucu: Şekil 4. 4'te gösterilen (a) S-parametreleri ve (b) grup gecikmesi.



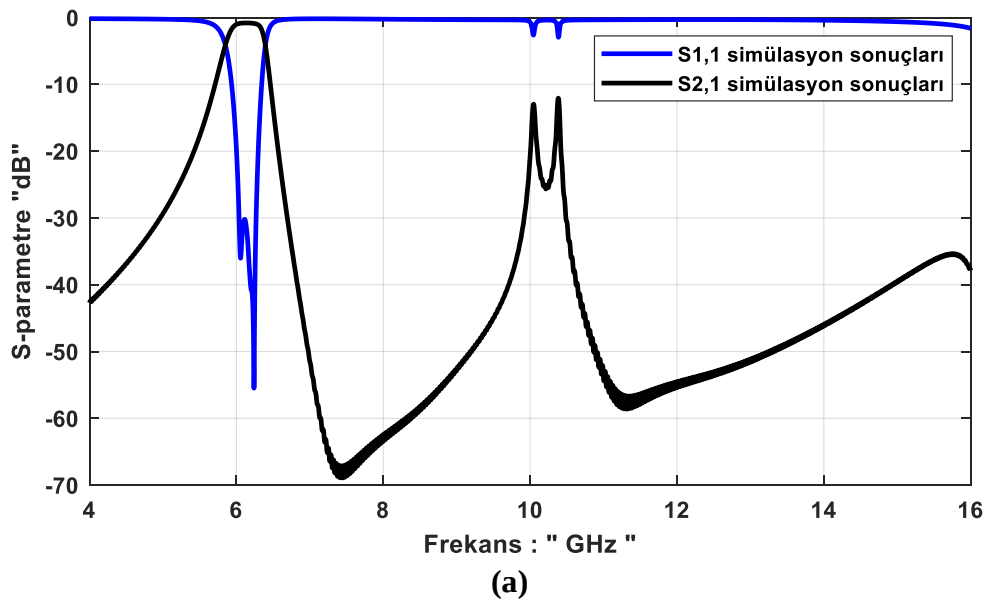


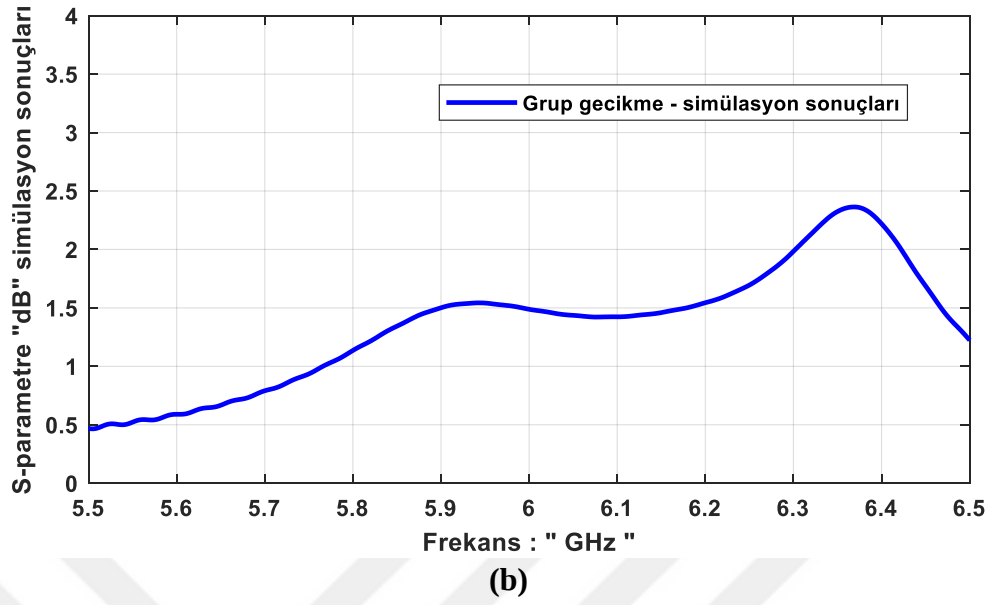
(b)

Şekil 4. 3. Üçüncü dereceden YGDK -ÇTY C-Bandı BGS; **a)** Üst düzlem; **b)** Alt düzlem; “Sarı bakır ve gri alt tabaka yalıtım malzemesini gösterir”

Çizelge 4. 2. C-Bandı filtre boyutları

Değişken	Uzunluk (mm)	Değişken	Uzunluk (mm)
Wm	1.519	SP	0.8
Lm	3.8	S	0.8
Wt	7.587	L	2.5
Lt	4.7	W	8.47
Wsiw	16.25	g1	0.5
d	1	g2	0.6
P	1.9	SubTh	0.508





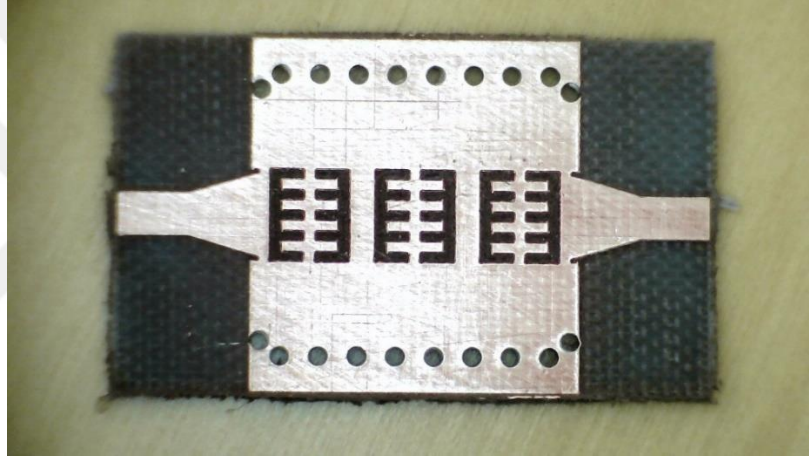
Şekil 4. 4. Üçüncü dereceden YGDK-ÇTY C-Bandı BGS; **a)** S-parametreleri; **b)** Grup gecikmesi

Tasarımda kullanılan YGDK ve mikroşerit arasındaki ara yüz geçişi, tam dalga simülatörü kullanılarak tasarlanmış bir konik geçiştir, filtre, 3 kademeli ÇTY hücresi ile yüklü YGDK 'nin basit bir kombinasyonudur. S-parametrelerinden, filtrenin 500 MHz'lık (5.925-6.425) GHz aralığında 3-dB bant genişliğine sahip olduğunu görebileceğimiz gibi, 6.175 GHz merkez frekansında 0.80 dB civarında ekleme kaybı (S_{21}) ve geri dönüş kaybı (S_{11}) geçiş bandında 30 dB'den daha iyi, bant dışı reddetme ile ilgili olarak, 6.425 GHz'de üst 3-dB kenar frekansından 7.4 GHz'de -68 dB sıfır iletimine keskin bir geçiş olduğunu görüyoruz, sahte tepki ortaya çıkıyor 10.05 ve 10.38 GHz'de -12.2 dB'den düşük. Filtrenin boyutuna gelince, filtre yekparelik özelliğini toplam $20 * 37.9$ mm² boyutunda korur. Daha detay Ek-1 (sayfa 41-44)'te sunulmuştur.

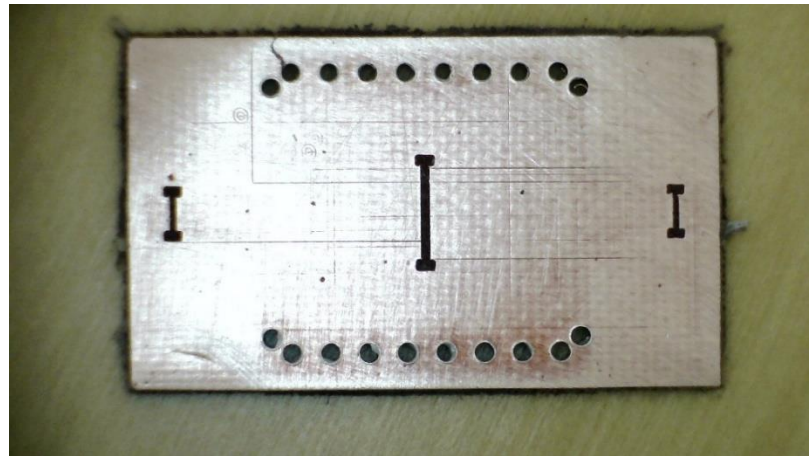
5. ÖLÇÜMLER

Bu bölümde fabrikasyon örnek ve ilgili ölçümleri tartışacağız. Fabrikasyon filtrelerin ölçümlerini yapmak için Ceyear 3672c-S Vektör Şebeke Analizörü kullanılmıştır. VNA kalibrasyonu, tam 2 portlu SOLT kalibrasyonu kullanılarak yapıldı. Aşağıdaki alt bölümlerde ölçüm sürecini ve ölçümler sırasında karşılaştığımız sorunları açıklıyoruz.

Üretilen X-bandı filtresinin örneği Şekil 5.1'de gösterilmektedir, imalat sırasında freze makinesi, yeterince sert olmamasına neden olan dokuma fiberglas takviyeli, PTFE bazlı kompozit substrat bazlı Rogers DiClad 880 plakalarında bir sorunla karşılaştı, esnekliğe ve bir dereceye kadar yumuşaklığa sahiptir, bu nedenle LPKF S104 freze makinesi, aşağıdaki Şekil 5.1'deki fabrikasyon numune ile fark edebileceğimiz gibi, işlem sırasında bakırın kenarlarında yüksek verimlilikte dağlama yapamaz.



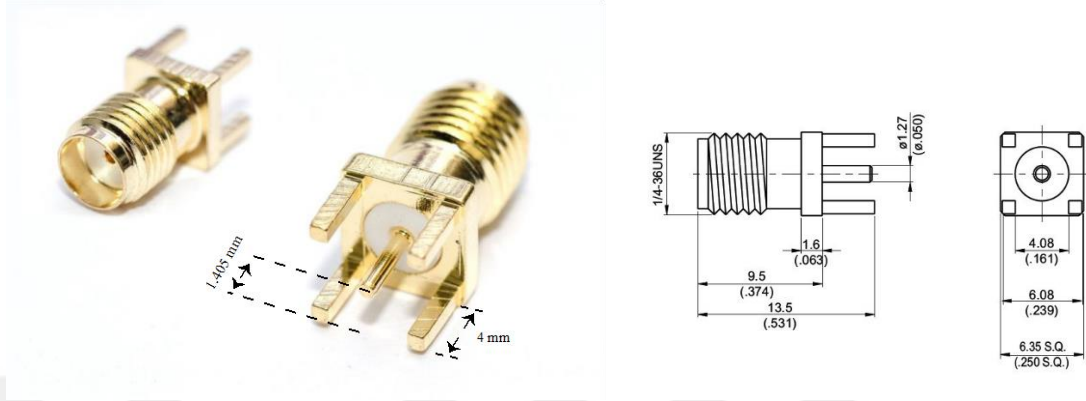
(a)



(b)

Şekil 5. 1. Fabrikasyon X-Band filtresi; **a)** Üst düzlem; **b)** Alt düzlem

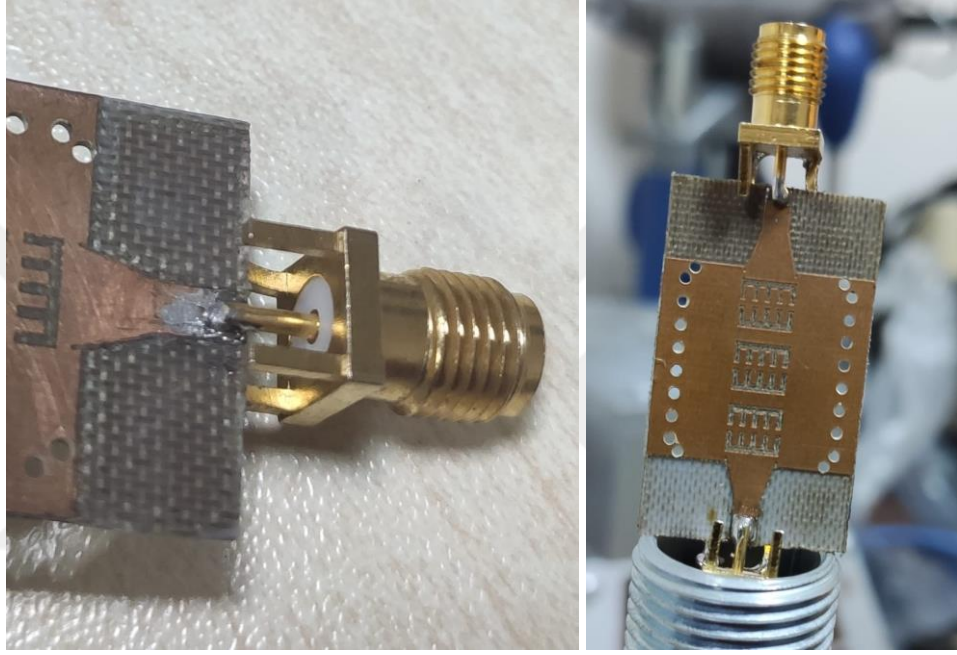
Ölçümler sırasında bizimle mevcut olan SMA konektörleri de Şekil 5.2'de gösterilmiştir. İç iletken pimi ile toprak ayakları arasındaki mesafe, 0.508 mm'lik plaka kalınlığına uymayan yaklaşık 1.4 mm, bu da kurulumda karışıklığa neden oldu ve böylece ölçülen sonuçlar.



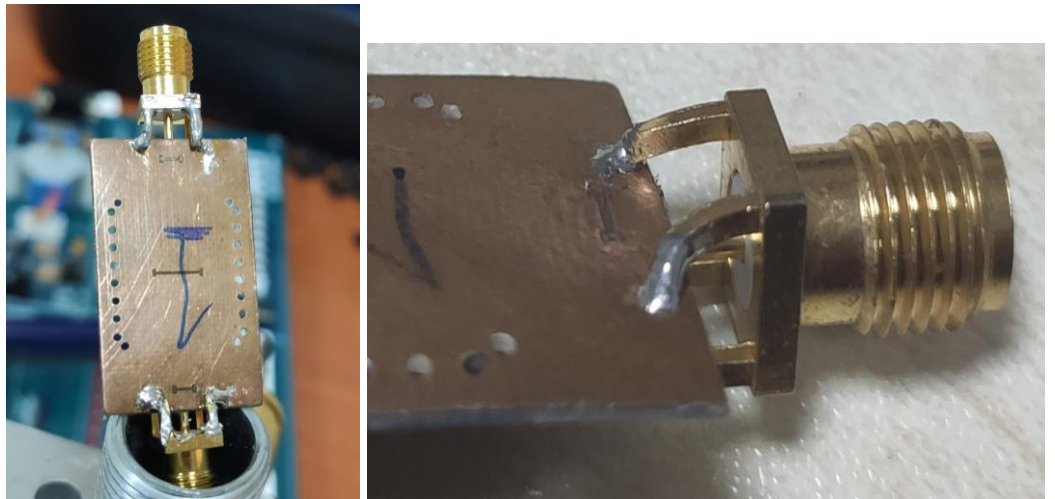
Şekil 5. 2. SMA konektör ve karşılık gelen boyutları

5.1. X-Bant ölçümleri

Şekil 5.3'te gösterilen SMA konektörünün ilk bağlatısından, SMA bacakları ve DGS iki başlı takoz hücreleri arasındaki elektromanyetik bağlantı nedeniyle ölçümleri etkileyebilecek herhangi bir rahatsızlığı önlemek için çalıştığımızı fark edebiliriz, böylece bacakları DGS iki başlı takoz hücrelerinden uzak tutmak için mümkün olduğunca kadar eğdik. Ayrıca, SMA iç pim uzunluğu filtrenin mikroşerit portu uzunluğundan daha büyüktür, bu yüzden konektörü SMA pimi mikroşerit portların ilk yarısına lehimledik Şekil 5.3'te gördüğümüz gibi.



(a)



(b)

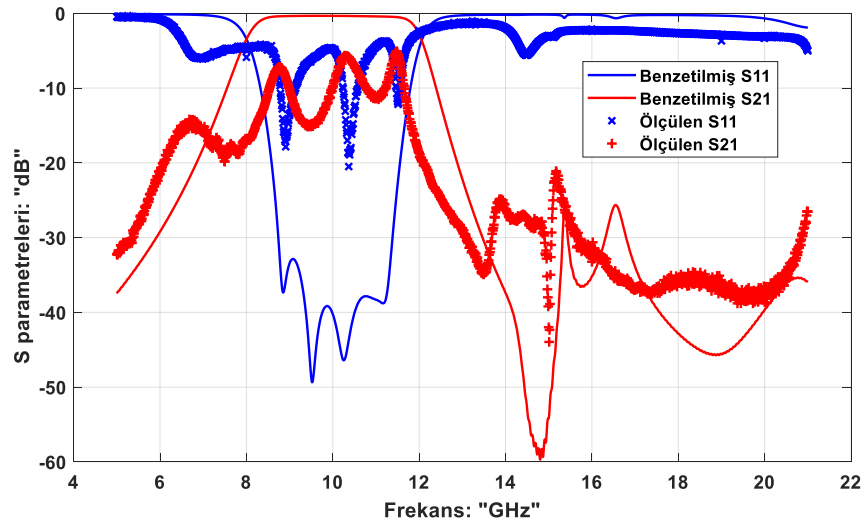
Şekil 5. 3. SMA konektörünün hatalı kurulumu; a) üst yüz; b) alt yüz

Konektörün yukarıda belirtildiği gibi takılmasından sonra, ölçümleri yapmak için filtreyi VNA uçlarıyla bağladık, Şekil 5.4'te gösterildiği gibi çok kötü sonuçlar aldık.

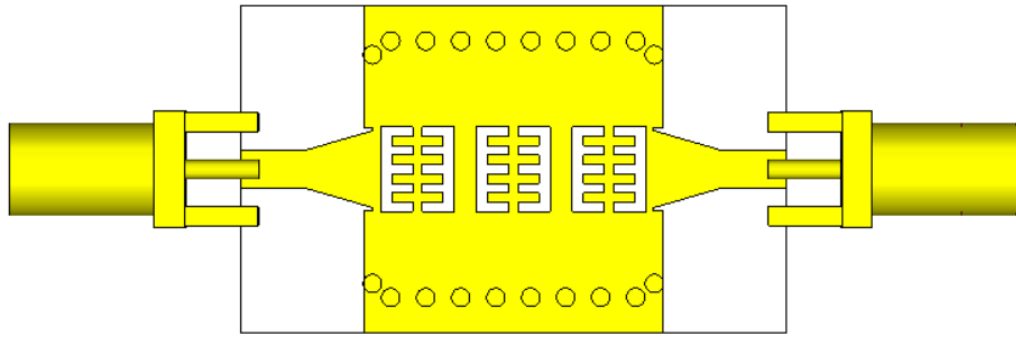


Şekil 5. 4. Ölçülen S_{21} yanıtı

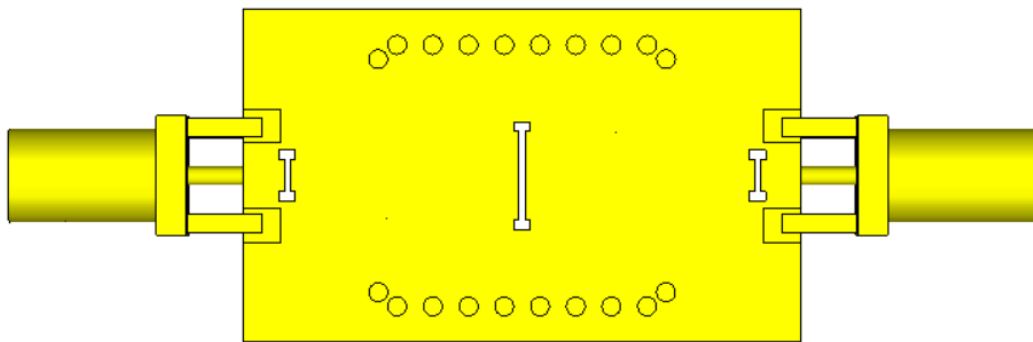
İdeal benzetim sonuçları ile ölçülen sonuçlar arasındaki bir karşılaştırma aşağıdaki Şekil 5.5'te gösterilmektedir, ölçülen sonuçların kabul edilemez olduğunu görebiliriz. Bu nedenle, ölçümlerin bozulmasına neden olan nedenleri araştırmak için, etkisini görmek için SMA konektörünü simülatördeki filtreyle modelledik, Şekil 5.6'da gösterildiği gibi, ölçümler sırasında kurduğumuz formda, SMA etkileri dikkate alınarak benzetim sonuçları ile Şekil 5.7'de gösterilen ölçülen sonuçlar arasında bir karşılaştırma.



Şekil 5. 5. Ölçüm sonuçlarına karşı ideal benzetim sonuçları



(a)

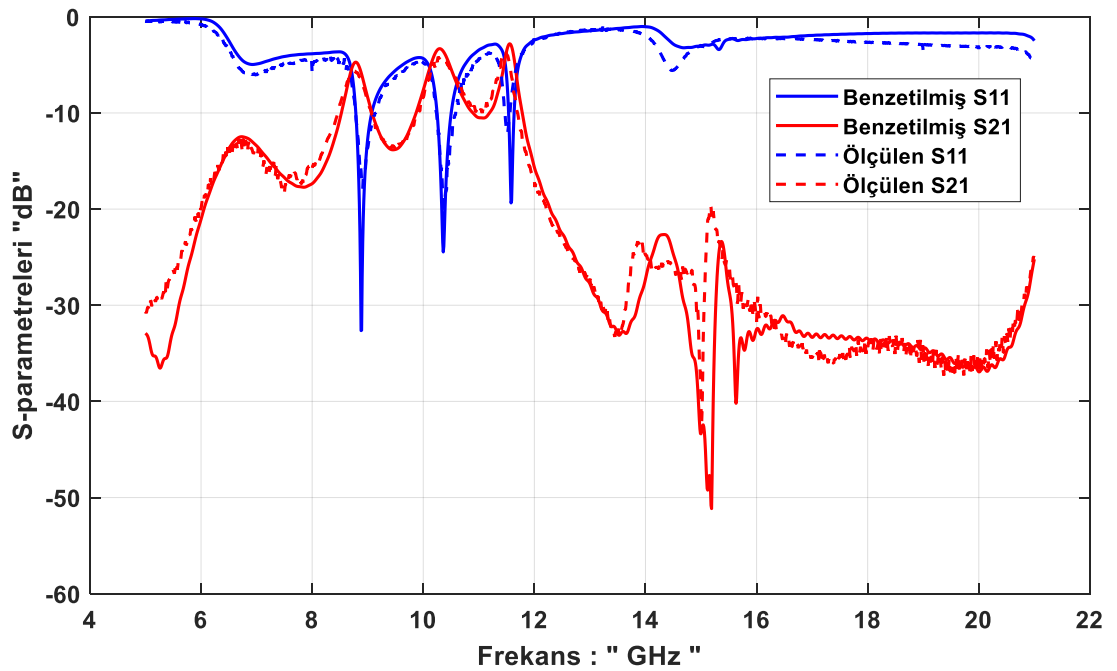


(b)



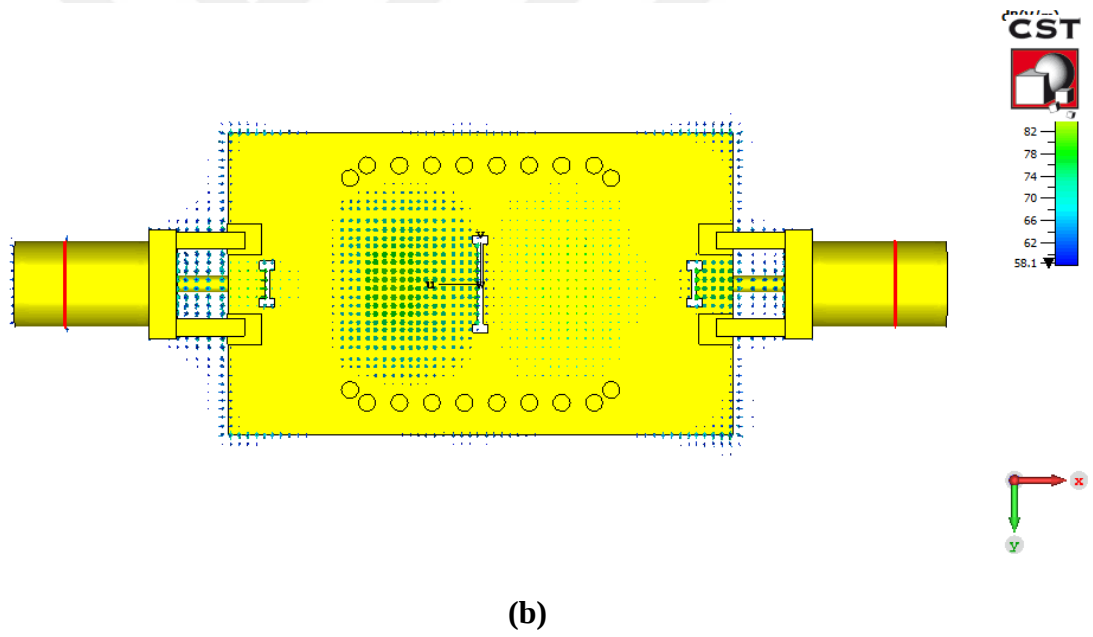
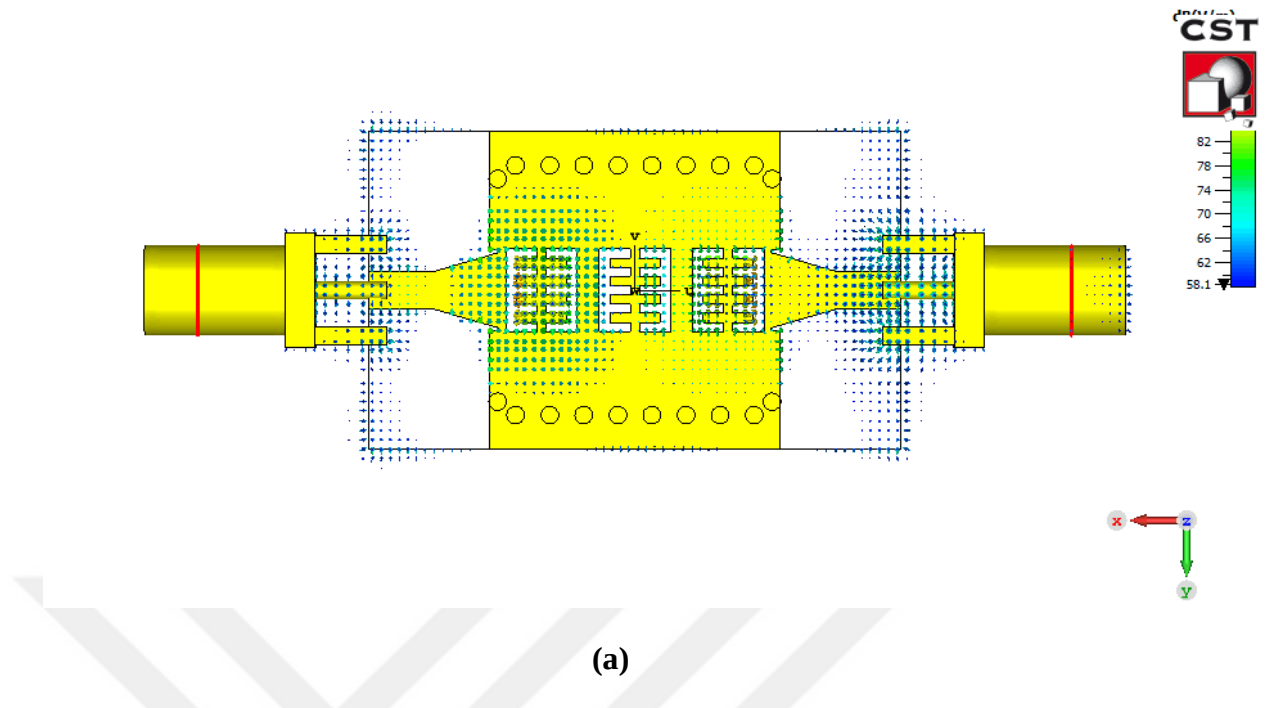
(c)

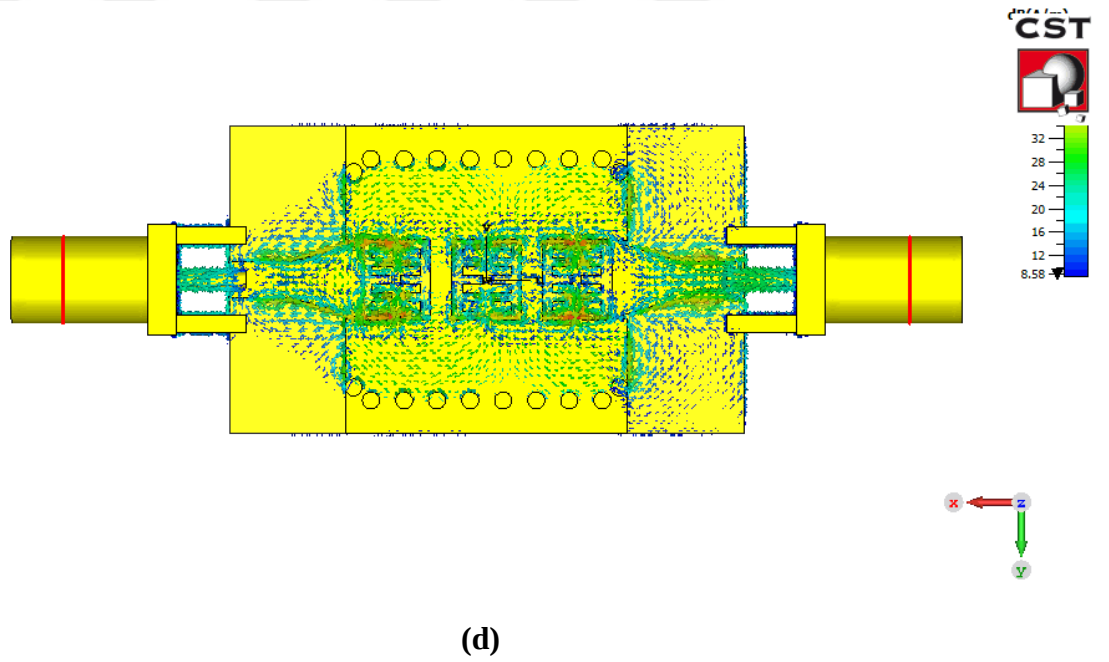
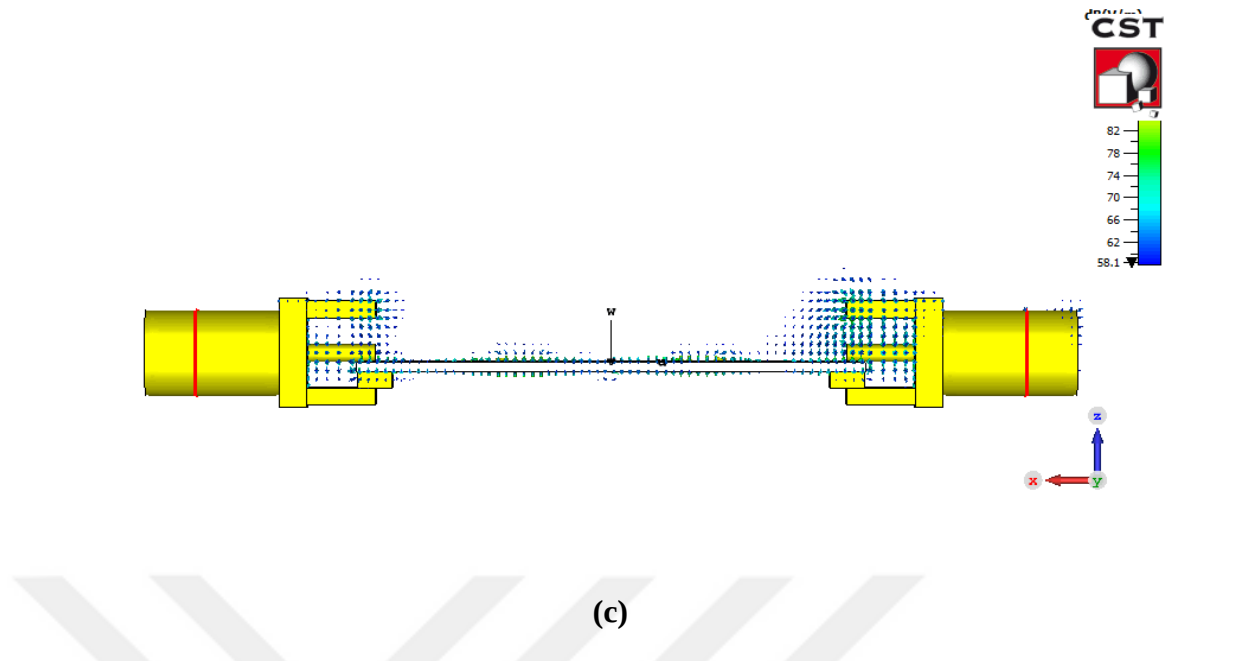
Şekil 5. 6. Benzetime SMA'nın dahil edilmesi; **a)** üst görünüm; **b)** alttan görünüm; **c)** yandan görünüm

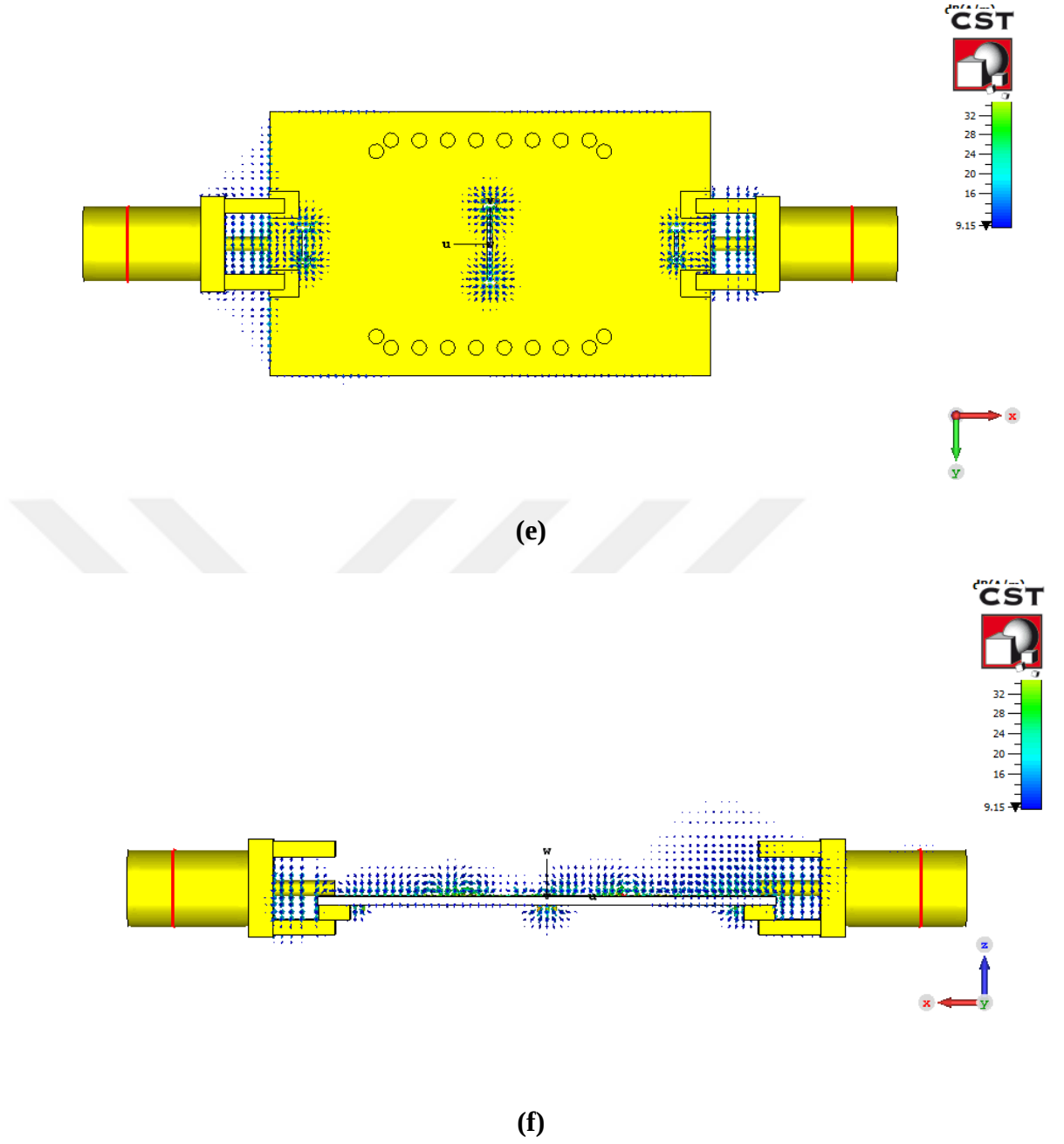


Şekil 5. 7. SMA hatalı kurulumu dikkate alınarak simüle edilmiş S-parametreleri, ölçülen S-parametrelerine karşı

Şekil 5.7'ye baktıktan sonra, bu kötü sonuçların, ölçüm kurulumunda kullanılan SMA konektörlerinin yanlış uç kurulumuyla ilişkilendirildiği sonucuna vardık. Ayrıca, Şekil 5.8'de gösterilen alan dağılımından, havaya düşen SMA'nın iç iletkeninin, harici bir indüktör olarak uç bağlantı elemanları ile filtrenin bağlantı noktaları arasındaki uyumu bozduğu fark edilebilir.



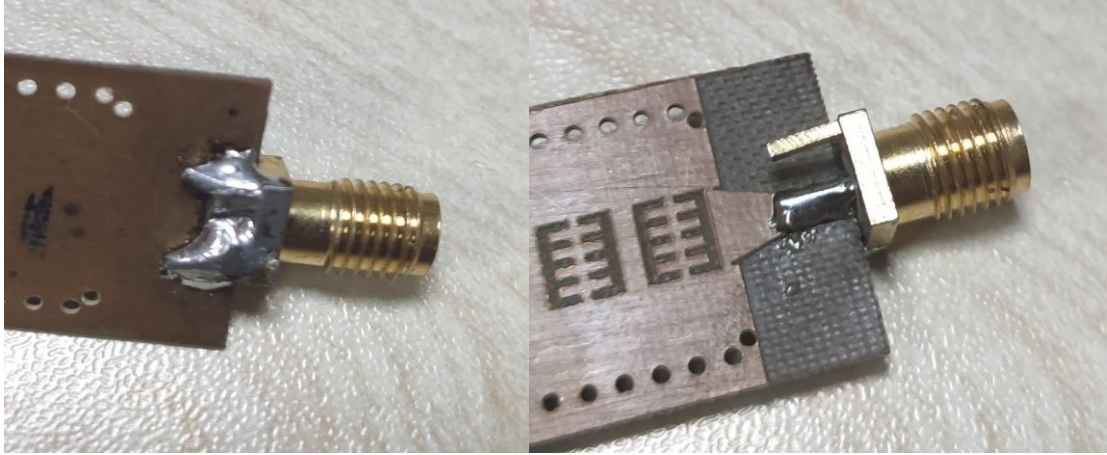




Şekil 5. 8. SMA yanlış kurulum nedeniyle alan dağılımı; **a)** E-alan üstten görünüşü; **b)** E-alanın alttan görünüşü; **c)** E-alan yandan görünüşü; **d)** H-alanı üstten görünüşü; **e)** H alanı alttan görünümü; **f)** H alanı yandan görünüşü;

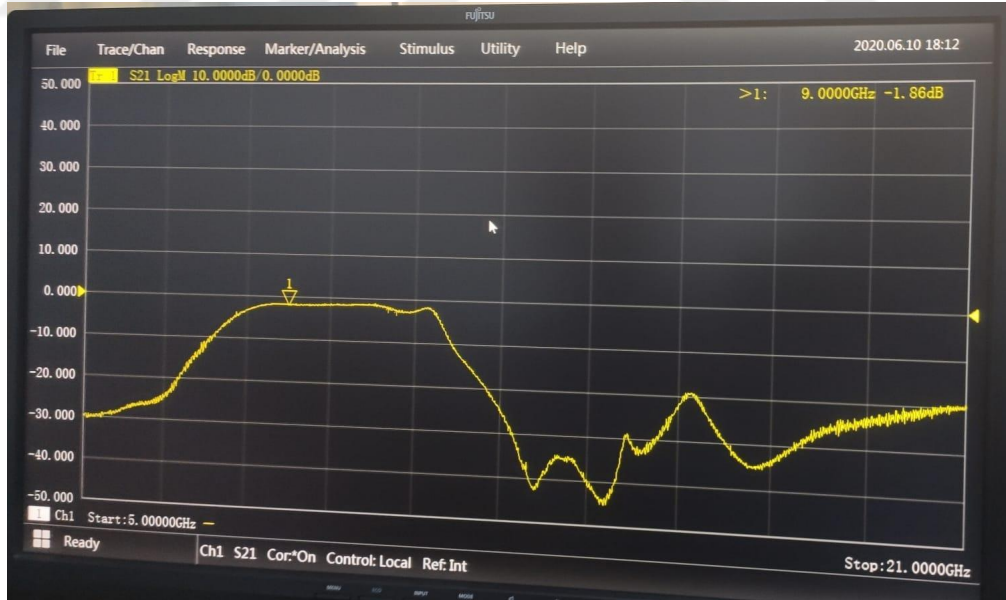
Bu nedenle, SMA toprak ayakları ve DGS iki başlı takozları arasındaki herhangi bir bağlantıya karşı herhangi bir problemten kaçınmaya çalışırken, filtre ve konektörler arasındaki eşleşmenin bozulmasına yol açtı. Bu nedenle, SMA kurulumunu düzelttik ve lehimlemeyi Şekil 5.9'da gösterildiği gibi güçlendirdik, burada SMA iç konektörleri ilgili

uçlara tamamen lehimlenerek bağlanmış ve ölçümleri etkileyen herhangi bir bağlantıyı önlemek için toprak ayakları iyice lehimlenmiştir.

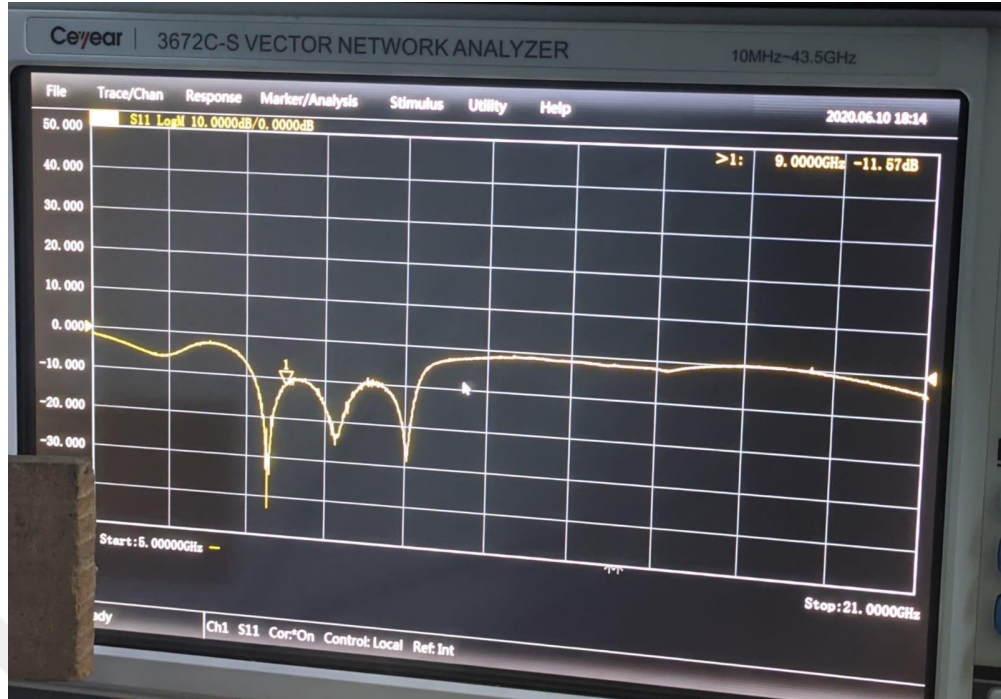


Şekil 5. 9. SMA konektör kurulumunu düzeltme

SMA kurulumunun düzeltilmesinden sonra ölçümleri gerçekleştirdik ve Şekil 5.10'da daha iyi sonuçlar elde ettik. SMA kurulumu düzeltmesinden sonra ölçülen S-parametre yanıtı ile ideal benzetimi yapılmış sonuçlar arasındaki karşılaştırma Şekil 5.11'de gösterilmektedir.

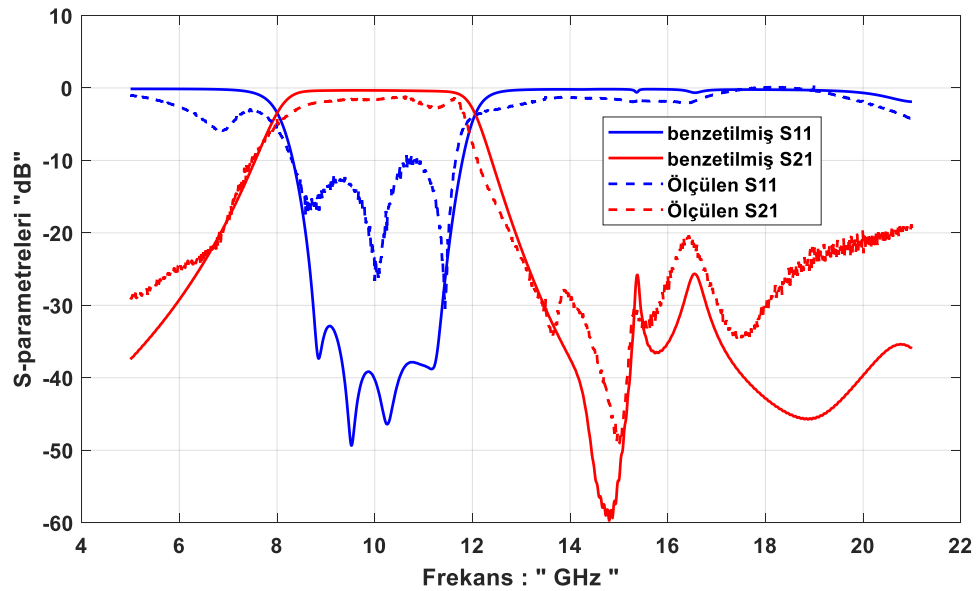


(a)



(b)

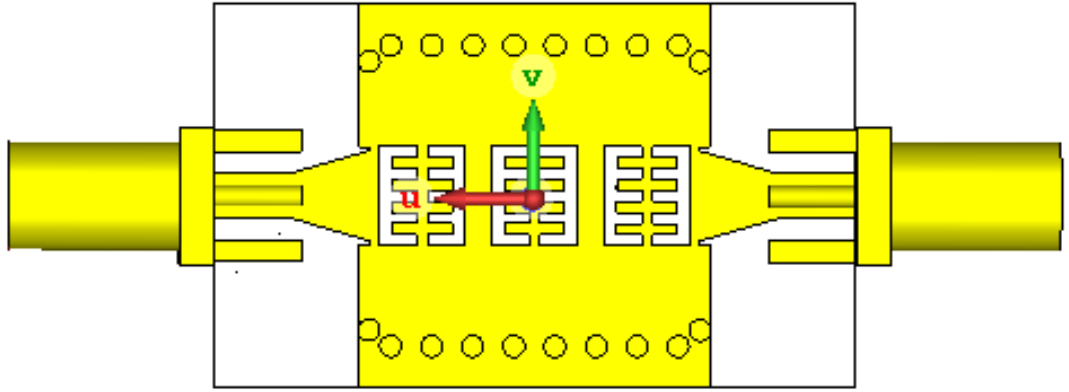
Şekil 5. 10. SMA kurulumunun düzeltilmesinden sonra ölçülen sonuçlar; a) ölçülen S_{21} yanıtı; b) ölçülen S_{11}



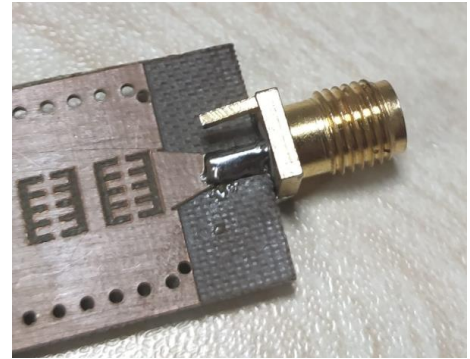
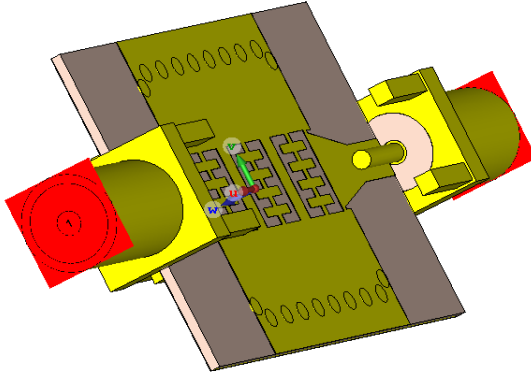
Şekil 5. 11. Ölçülen S-parametrelerine karşı ideal simülasyon

Ölçülen iletim parametresi (S_{21}), merkez frekansı ve bant dışı reddetme açısından benzetimi yapılan (S_{21}) ile uyumludur, ancak araya girme kaybı farkı vardır.

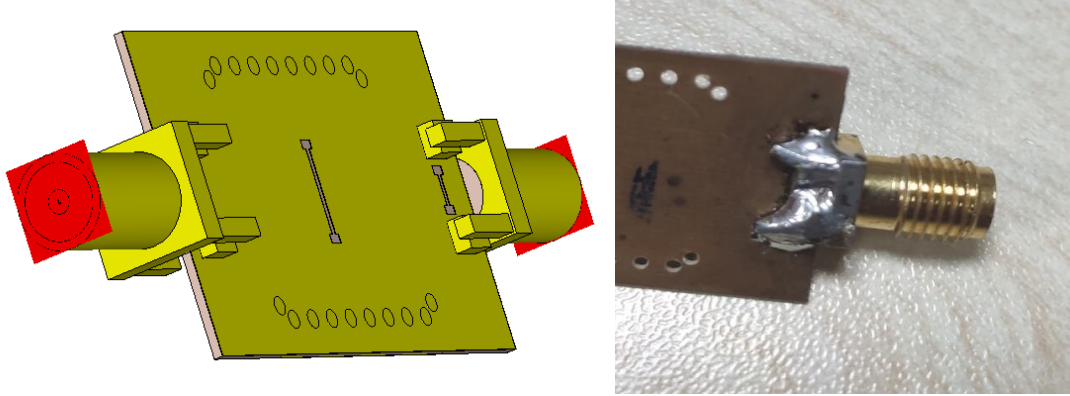
Burada 10 GHz merkez frekansında benzetimde görülen araya girme kaybı yaklaşık 0.33 dB olurken ölçülen cevap yaklaşık 1.6 dB'dir. Ayrıca alt ve üst frekans sınırlarında geçişin kenarlarında hafif bir sapma vardır. Şekil 5.11'den gördüğümüz gibi geri dönüş kaybı açısından, benzetim ile çalışma bandı içindeki ölçümler arasında bir fark vardır. Bu nedenle, bu konulara açıklama getirmek için, benzetimin SMA konektörünün neredeyse ölçümlerde olduğu gibi dahil edilmesini dikkate alarak yürütülmesi için benzetim programına geri dönüyoruz. Şekil 5.12'de SMA konektörlerinin de dahil edildiği bir benzetim şematiği gösterilmektedir. Şekil 5.13'te SMA konektörlerini dikkate alan benzetim sonuçları ile ölçüm sonuçları birlikte gösterilmektedir.



(a)

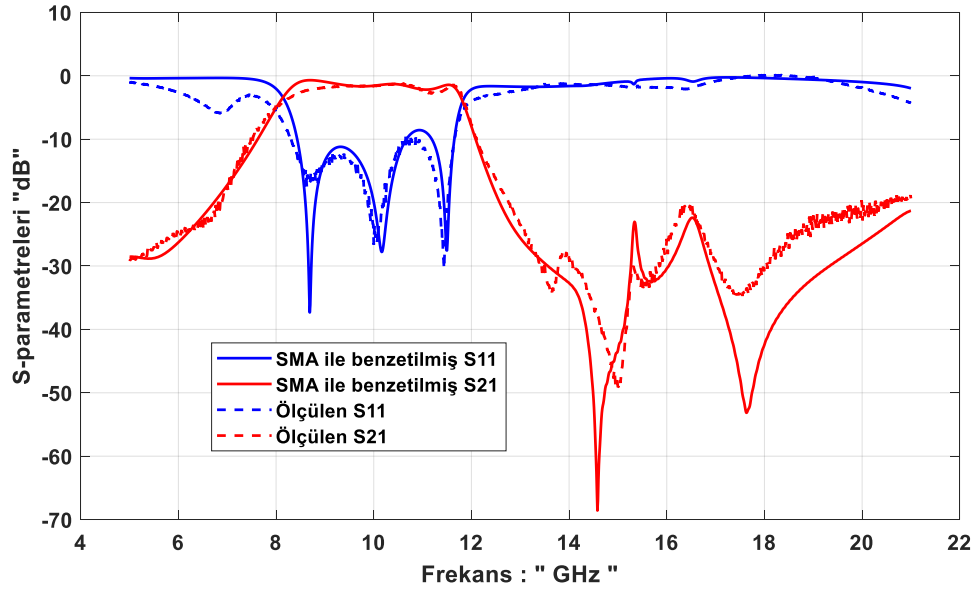


(b)



(c)

Şekil 5. 12. Simülatöre SMA konektörleri dahil etmek; **a)** üstten görünüm; **b)** üst yüzün eğik görünümü; **c)** alt yüzün eğik görünümü



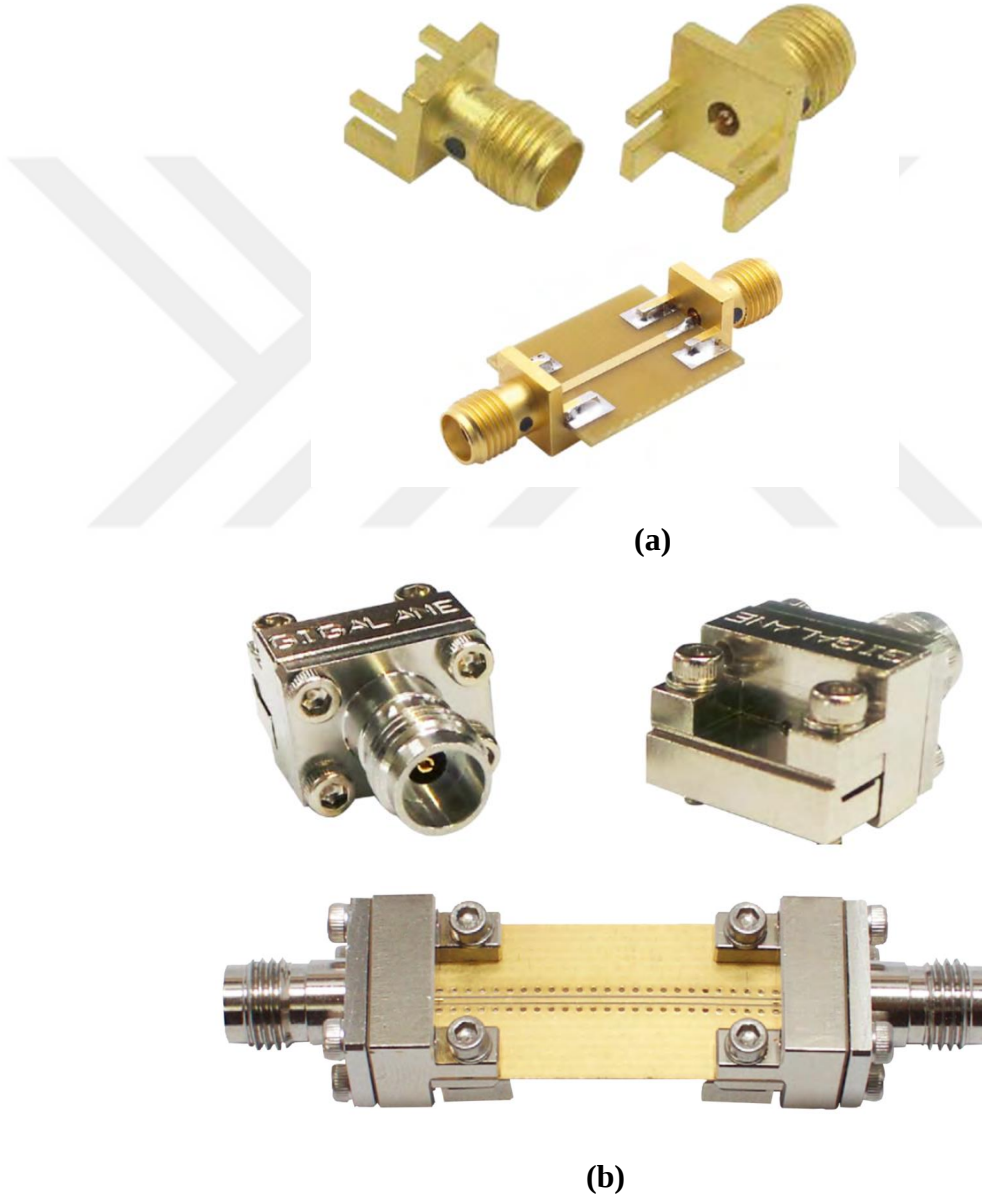
Şekil 5. 13. Benzetim SMA-dahil, Ölçülen sonuçlar karşı

Burada ölçülen sonuçlar, merkezi frekans, bant genişliği, geri dönüş kaybı ve yerleştirme kaybı ve aynı zamanda bant dışı reddetme açısından SMA'nın dahil edilmesinden sonraki benzetim ile iyi bir uyum içindedir, yaklaşık -1.6 dB'lik bir ölçüm kaybı var, ayrıca, -10 dB'den daha iyi ölçülen bir geri dönüş kaybı var, SMA konektörünün varlığını dikkate alan benzetim sonuçlarıyla iyi anlaşılır.

Dolayısıyla, Şekil 5.13'e baktıktan sonra, Şekil 5.11'de gösterilen ideal benzetim sonuçları ile ölçüm sonuçları arasındaki farkın, ölçüm kurulumunda kullanılan ancak

tasarım sırasında dikkate alınmayan SMA konektörlerine atfedildiği sonucuna vardık. Bu konektörleri ekleme kaybını ve geri dönüş kaybını etkiler.

Ayrıca, daha ince ve daha kısa bir iç pim konektörüne sahip olan ve iç iletken ile plaka kalınlığına uygun toprak ayakları arasında ayrı bir mesafeye sahip yüksek performanslı konektörler kullanılmalıdır. Veya iyi oturan ve SMA konektörlerinden daha yüksek frekanslara kadar çalışan K-konektörü kullanarak, bu konektörler Şekil 5.14'te gösterilmiştir, bu tip konektörler mükemmel dönüş Kaybı ve 26 GHz'den daha yüksek frekanslara kadar daha iyi ekleme kayıpları sağlar.



Şekil 5. 14. Yüksek performanslı konektörler; **a)** yüksek performanslı SMA konektörü 26,5 GHz'e kadar hizmet verir; **b)** 2,92 mm veya k- konektörü 50 GHz'e kadar hizmet eder

6. SONUÇLAR

Bu çalışmada, diğer düzlemsel yapılara kolayca entegre olma özelliğine sahip olan Yama Giydirilmiş Dalga Kılavuzu (YGDK) ve Çentikli Taban Yapılar (ÇTY) tekniklerine dayanan iki bant geçiren filtrenin tasarım artılarını inceledik. İlk BPF, X-bandı uygulamaları için geniş bantlı bir filtredir ve ikincisi, C-bandı Uydu Haberleşmene adanmış geniş bantlı bir filtredir. Filtrelerin modelleme ve benzetim süreçleri CST Mikrodalga Stüdyosu'nda 3D modeller yapılarak gerçekleştirildi.

X-bandı önerilen filtre, düşük bir ekleme kaybı, keskin ve geniş bant dışı bastırma ile iyi bir geri dönüş kaybına sahiptir. Performans geliştirme sürecini içeren tasarım aşamalarını sergiledik. YGDK- ÇTY hücreleri kullanılarak oluşturulan filtre prototipi ve geçiş bandında iyi bir performansla sahipti, ancak durdurma bandında sahte tepkiden mustarip. Durdurma bandının bant genişliğini genişletmek ve sahte tepkilerin bastırılmasını arttırmak amacıyla, daha iyi geçiş bandı performansı ile birlikte çok geniş bir durdurma bandı reddi ve azaltılmış sahte tepki ile sonuçlanan YGDK'nın zemin düzleminde iki başlı takoz yuvaları kazanmıştır.

C-bandı filtresi C-bandı uydu iletişim uygulamaları için tasarlanmıştır, esas olarak düzlemsel olmayan dalga kılavuzu teknolojisinin akran filtresinin benzetimi için tasarlanmıştır. Elde edilen sonuçlar görece çok ucuz olan bu teknolojinin kendinden daha pahalı hantal bir teknoloji ile karşılaştırıldığında kabul edilebilir başarılı sonuçlar sunduğu görülmektedir. Ayrıca, önerilen kompakt boyutlara sahip filtreler geçiş bantlarında neredeyse düz bir grup gecikmesi sunar.

Bu tezdeki çalışmanın sonuçları, yüksek performanslı geniş bant filtrelerin tasarlanmasında bu teknolojinin geçerliliğine ve yeteneğine ışık tutmaktadır. Daha fazla boyut küçültme Yarım Mod Yama Giydirilmiş Dalga Kılavuzu kullanılarak yapılabilir. İris tipi rezonatörlere dayanan filtrelerin tasarımında kuplaj matris sentezinin kullanılması veya daha yüksek Q faktörüne sahip yeni rezonatörlerin aranması gibi yeni dar bantlı filtrelerin tasarımında YGDK ile birlikte başka yöntemler de uygulanabilir, bu filtreler ÇTY ile desteklenebilir veya durdurma bandı performansını daha da artırmak için başka bir iletim sıfırları eklemek üzere giriş ve çıkış bağlantı noktalarında mahmuz hatları kullanılarak desteklenebilir.

7. KAYNAKLAR

- Hirokawa, J., & Ando, M. 1998. Single-layer feed waveguide consisting of posts for plane TEM wave excitation in parallel plates. **IEEE Transactions on Antennas and Propagation**, 46(5): 625-630.
- Deslandes, D., & Wu, K. 2001. Integrated microstrip and rectangular waveguide in planar form. **IEEE Microwave and Wireless Components Letters**, 11(2): 68-70.
- L. H. Weng, Y. C. G., X. W. Shi, and X. Q. Chen. 2008. AN OVERVIEW ON DEFECTED GROUND STRUCTURE. **Progress In Electromagnetics Research B**, 7: 173-189.
- Y. Huang, Z. S., and L. Liu. 2013. A Substrate Integrated Waveguide Bandpass Filter Using Novel Defected Ground Structure Shape. **Progress in Electromagnetics Research-Pier**, 135: 201-213.
- Steer, M. B. S. a. M. B. 2016. **FOUNDATIONS FOR MICROSTRIP CIRCUIT DESIGN**: John Wiley & Sons, Ltd.
- HONG, J.-S. 2011. **Microstrip Filters for RF/Microwave Applications**. New York ;Chichester: Wiley.
- Pozar, D. M. 2012. **Microwave Engineering**. University ofMassachusetts at Amherst: John Wiley & Sons, Inc.
- Ke, W., Deslandes, D., & Cassivi, Y. 2003. **The substrate integrated circuits - a new concept for high-frequency electronics and optoelectronics**. Paper presented at the 6th International Conference on Telecommunications in Modern Satellite, Cable and Broadcasting Service, 2003. TELSIS 2003.
- Deslandes, D., & Wu, K. 2001. Integrated microstrip and rectangular waveguide in planar form. **IEEE Microwave and Wireless Components Letters**, 11(2): 68-70.
- Caleffo, R. C. 2016. New Design Procedure to Determine the Taper Transition for Impedance Matching Between Microstrip Line and SIW Component. **Journal of Microwaves, Optoelectronics and Electromagnetic Applications**, 15: 247-260.
- Wang, J., Hao, Z., & Kui-Kui, F. 2016. **A 110–150 GHz SIW-rectangular waveguide transition for terahertz applications**. Paper presented at the 2016 IEEE MTT-S International Microwave Workshop Series on Advanced Materials and Processes for RF and THz Applications (IMWS-AMP).
- Deslandes, D., & Ke, W. 2006. Accurate modeling, wave mechanisms, and design considerations of a substrate integrated waveguide. **IEEE Transactions on Microwave Theory and Techniques**, 54(6): 2516-2526.
- Y. Cassivi, L. P., P. Arcioni, M. Bressan, K. Wu, and G. Conciauro. 2002. Dispersion Characteristics of Substrate Integrated Rectangular Waveguide **IEEE MICROWAVE AND WIRELESS COMPONENTS LETTERS**, 12.

- Kordiboroujeni, Z., & Bornemann, J. 2014. New Wideband Transition From Microstrip Line to Substrate Integrated Waveguide. *IEEE Transactions on Microwave Theory and Techniques*, 62(12): 2983-2989.
- Jong-Im, P., Chul-Soo, K., Juno, K., Jun-Seok, P., Yongxi, Q., Dal, A., & Itoh, T. 1999. *Modeling of a photonic bandgap and its application for the low-pass filter design*. Paper presented at the 1999 Asia Pacific Microwave Conference. APMC'99. Microwaves Enter the 21st Century. Conference Proceedings (Cat. No.99TH8473).
- Ahn, D., Park, J., Kim, C., Kim, J., Qian, Y., & Itoh, T. 2001. A design of the low-pass filter using the novel microstrip defected ground structure. *IEEE Transactions on Microwave Theory and Techniques*, 49(1): 86-93.
- Z. He, J. C., Z. Shao, X. Li, and Y. Huang. 2013. A Novel Power Divider Integrated with SIW and ÇTY Technology. *Progress In Electromagnetics Research*, 139: 289-301.

8. EKLER**8.1. EK-1 Tezin İngilizce versiyonu****T.C.****AKDENİZ ÜNİVERSİTESİ****YAMA GİYDİRİLMİŞ DALGA KILAVUZU (YGDK)
TEKNOLOJİSİ İLE X BANDI
BANT GEÇİREN FİLTRE TASARIMI****Mohammed R. A. NASSER****FEN BİLİMLERİ ENSTİTÜSÜ****ELEKTRİK ELEKTRONİK MÜHENDİSLİĞİ****ANABİLİM DALI****YÜKSEK LİSANS TEZİ****HAZİRAN 2020****ANTALYA**

**T.C.
AKDENİZ ÜNİVERSİTESİ**



**FILTER DESIGN USING THE SUBSTRATE INTEGRATED
WAVEGUIDE (SIW) TECHNOLOGY
ON X BAND**

Mohammed R. A. NASSER

INSTITUTE OF SCIENCE

ELECTRICAL & ELECTRONICS ENGINEERING

DEPARTMENT

MASTER'S THESIS

**JUNE 2020
ANTALYA**

**T.C.
AKDENİZ ÜNİVERSİTESİ**



**FILTER DESIGN USING THE SUBSTRATE INTEGRATED
WAVEGUIDE (SIW) TECHNOLOGY
ON X BAND**

Mohammed R. A. NASSER

**INSTITUTE OF SCIENCE
ELECTRICAL & ELECTRONICS ENGINEERING
DEPARTMENT
MASTER'S THESIS**

JUNE 2020

ANTALYA

**T.C.
AKDENIZ UNIVERSITY
INSTITUTE OF SCIENCE**

**FILTER DESIGN USING THE SUBSTRATE INTEGRATED
WAVEGUIDE (SIW) TECHNOLOGY
ON X BAND**

**Mohammed R. A. NASSER
ELECTRICAL & ELECTRONICS ENGINEERING
DEPARTMENT
MASTER'S THESIS**

**This thesis was supported by Akdeniz University Scientific Research Projects
Coordination Unit with the project numbered FYL-2020-5103.**

JUNE 2020

**T.C.
AKDENİZ UNIVERSITY
INSTITUTE OF SCIENCE**

**FILTER DESIGN USING THE SUBSTRATE INTEGRATED
WAVEGUIDE (SIW) TECHNOLOGY
ON X BAND**

**Mohammed R. A. NASSER
ELECTRICAL & ELECTRONICS ENGINEERING
DEPARTMENT
MASTER'S THESIS**

This thesis was accepted by the jury committee of the 19/06/2020

Prof. Dr. Selçuk HELHEL (Supervisor)

Prof. Dr. Şükrü ÖZEN

Assistant Prof. Dr. I. Bahadır BAŞYİĞİT

ABSTRACT**FILTER DESIGN USING THE SUBSTRATE INTEGRATED
WAVEGUIDE (SIW) TECHNOLOGY
ON X BAND****Mohammed R. A. NASSER****MSc Thesis in Electrical & Electronics Engineering****Supervisor: Prof. Dr. SELÇUK HELHEL****June 2020; 65 pages**

Planer band-pass filters are considered as a major part of any RF/Microwave device. The increasing demands on communications and radar systems require filters that provide all of compactness and lightweight with superior performance and easy to integrate with other circuits. In this thesis, we look at new planer technology called substrate integrated waveguides (SIW), developed recently with characteristics mixes the affirmative characteristics of the metallic waveguides and microstrip technologies, has opened the horizons in microwave filter design to provide new designs to match the high requirements of the increased demand.

In this thesis, we have designed two 3rd order compact filters using the Substrate integrated waveguide (SIW) technology and the Defected ground structure (DGS). Both of the designs are using the SIW-microstrip taper transition to ensure wideband adaption with microstrip technology.

The first filter is a Wideband bandpass filter designed for X-band application has a center frequency at 10 GHz, with fractional bandwidth (FBW 40%), the structure composed by SIW and two types of defected ground structures (DGS) loaded in both of the Top and bottom plane of the SIW and its characterized by the wide free of spurious upper stopband, while the second filter is a broadband filter designed for C-band Space communications has a center frequency at 6.175 GHz, with 500 MHz bandwidth (FBW=8.3%), the structure composed by SIW and 3-DGS cells on the top plane of the SIW.

The filters were built on a single-layer Rogers DiClad 880 substrate, and they were designed and optimized using CST MW studio the time domain solver works by using the "Finite-difference time-domain method (FDTD) numerical method" technique.

KEYWORDS: Band-pass filters, substrate-integrated waveguide (SIW), defected ground structures (DGS), wideband taper transition, X-band, C-band.

COMMITTEE:

Prof. Dr. Selçuk HELHEL (Supervisor)

Prof. Dr. Şükrü ÖZEN

Assistant Prof. Dr. I. Bahadır BAŞYİĞİT

PREFACE

Compact size, low insertion loss, good return loss as well as smooth group delay. All of these filters related requirements cannot be satisfied all at once by the traditional technologies, as known the high performance is related always to non-planer waveguide technology which has a bulky size. As for the Planer technologies features by the compactness and ease of integrability with other circuits but these pros come at the expense of fragile performance. So, in this thesis, we discuss the new technology called Substrate Integrated Waveguide (SIW) that helped in eliminating the fragile performance of the planer technologies and provide new principles on filter design.

I would like to express my sincere gratitude to my supervisor Professor Prof. Dr. Selçuk HELHEL for accepting me as his student and for his support of my research and to express my thanks to Prof. Bektaş ÇOLAK for helping during the measurements.

I would like to extend my sincerest thanks to the Turkish scholarship (YTB) for granting me financial support, that paved my way in realizing my dream which would have not been com to reality without their help. Also, my great thanks to Akdeniz University Scientific Project Support Unit (BAP) for supporting my research. Also, I would like to thank Akdeniz University EMUMAM directorate that all facilities used in this study were granted by the State Planning Organization - Turkey (Grant Number: 2007K120530-DPT).

Finally, I would like to especially thank my mother my family, and friends for their invaluable support and constant encouragement.

TABLE OF CONTENTS

ABSTRACT	i
PREFACE	ii
ACADEMIC STATEMENT	v
SYMBOLS AND ABBREVIATIONS	vi
LIST OF FIGURES	ix
LIST OF TABLES	xiii
1. INTRODUCTION	1
1.1. Introduction	1
1.2. Problem Statement.....	2
1.3. Objectives	4
1.4. Scopes.....	5
1.5. Organization of the thesis.....	5
2. THEORETICAL BACKGROUND	6
2.1. Transmission Lines.....	6
2.1.1. Introduction.....	6
2.1.2. Microstrip lines (HONG, 2011).....	6
2.1.3. Rectangular Waveguides	9
2.1.4. Substrate integrated waveguides.....	13
2.2. Microstrip – SIW Transition	19
2.3. Defected Ground Structures	21
2.3.1. DGS Applications in Microwave Circuit (L. H. Weng, 2008).....	26
3. METHODOLOGY	27
4. DESIGN PROCESS AND RESULTS.....	30
4.1. X– Band Filter	30
4.1.1. SIW & Taper transition design	30
4.1.2. DGS cell design	33
4.1.3. 1 st -Order SIW-DGS bandpass filter.....	35
4.1.4. SIW-DGS 3 rd order BPF	37
4.1.5. SIW-DGS 3 rd BPF – with one Dumbbell cell.....	39
4.1.6. SIW-DGS BPF - 3 Dumbbell DGSs etched in Ground plane	41
4.2. C– Band Filter	44

5. MEASUREMENTS	48
5.1. X-Band measurements.....	50
5.2. Discussing the Reasons for the Results Variation	62
6. CONCLUSION	63
7. RESOURCES.....	64



ACADEMIC STATEMENT

Master's thesis name "FILTER DESIGN USING THE SUBSTRATE INTEGRATED WAVEGUIDE (SIW) TECHNOLOGY ON X BAND" is written with academic rules and ethical values. I declare that I had references all of the information that does not belong to me.

19/06/2020

Mohammed R. A. NASSER

SYMBOLS AND ABBREVIATIONS

Symbols

Hz : Hertz (1/second)

MHz : Mega (10^6) hertz

GHz : Giga (10^9) hertz

ϵ_r : Relative permittivity or dielectric constant

$\tan \delta$: Dielectric loss tangent

W : Microstrip transmission line width

T : Laminate's cladding thickness

h : Substrate thickness

μ : Permeability

ϵ_{re} : Effective dielectric constant

Z_c : Characteristic impedance

C_0 : Velocity of light in free space

λ_0 : Free space wavelength

λ_g : Guided wavelength

θ : Electrical length

β : Propagation constant

f : Frequency

f_c : Cutoff frequency

σ : Conductivity

mm : Millimeter

mm^2 : Square millimeter, an international unit of area

dB : Decibel

ns : Nanoseconds

Abbreviations

RF	: Radio frequency
CAD	: Computer-aided design
EM	: Electromagnetic
SIW	: Substrate integrated waveguide
RWG	: Rectangular waveguide
Q-factor	: Quality factor
PCB	: printed circuit board
SIW	: Substrate integrated waveguide
DGS	: Defected ground structures
BPF	: Band-pass filter
C-Band	: Frequencies ranging from 4.0 to 8.0 gigahertz
X-Band	: Frequencies ranging from 8.0 to 12.0 gigahertz
CST	: Central Standard Time
3D	: Three-dimensional
FBW	: Fractional bandwidth
TL	: Transmission line
TEM	: Transverse electromagnetic
TE	: Transverse electric
TM	: Transverse magnetic
DC	: Direct current
LTCC	: Low-temperature cofired ceramic
PTFE	: Polytetrafluoroethylene
LC circuit	: Resonant circuit of an <u>inductor</u> “L”, and a <u>capacitor</u> “C”
S-parameter	: Scattering parameters
TM resonators	: Transverse magnetic-mode resonators

MW : Microwave

S_{21} : The forward transmission (from port 1 to port 2)

S_{11} : The input reflection coefficient with the output of the network terminated by a matched load

LPF : Low pass filter

HPF : High pass filter



LIST OF FIGURES

Figure 1. 1. Simplified RF front end transceiver block diagram.....	1
Figure 1. 2. Filter types according to frequency selectivity	1
Figure 1. 3. Rectangular waveguide samples	3
Figure 1. 4. Planer waveguides; a) Microstrip; b) coplanar waveguide	3
Figure 1. 5. Substrate integrated waveguide's general structure.....	4
Figure 2. 1. Microstrip line structure.....	6
Figure 2. 2. Microstrip quasi-TEM mode where the dashed lines represent the magnetic field lines and the solid lines represent the electric field lines.....	7
Figure 2. 3. Typical waveguides; a) circular waveguide; b) rectangular waveguide; c) twist; 90-degree elbow	9
Figure 2. 4. Waveguide cutoff frequency.....	9
Figure 2. 5. Geometry of a rectangular waveguide with perfectly conducting walls, filled with a lossless material	10
Figure 2. 6. Rectangular waveguide TE_{10} mode field distribution; a) Electric field lines (front view); b) Magnetic field (top view); c) Electric field (top view)	12
Figure 2. 7. Performance gap between planar and non-planar structures led to SIW development	14
Figure 2. 8. SIW and its equivalent RWG.....	14
Figure 2. 9. SIW transmission line with TE_{10} mode; a) cross-section; b) top view.....	15
Figure 2. 10. Top view of the SIW.....	16
Figure 2. 11. Effect of changing the SIW's width (a) on the TE_{10} mode cutoff frequency	18
Figure 2. 12. SIW excitation using microstrip port; a) TE_{10} -SIW; b) <i>Quasi – TEM</i> microstrip line; c) Electric field distribution in SIW transmission line exploiting the TE_{10} mode excitation using Quasi-TEM mode	19
Figure 2. 13. Configuration of the microstrip-SIW transition.....	20
Figure 2. 14. Structural parameters of the new taper-via transition between microstrip and SIW (Kordiboroujeni & Bornemann, 2014).....	20
Figure 2. 15. Simple DGS slots.....	22

Figure 2. 16. Microstrip line loaded by dumbbell DGS cells (Jong-Im et al., 1999); a) LPF structure; b) microstrip transmission line; c) dumbbell DGS cells etched in the ground plane; d) <i>S</i> -parameters response	23
Figure 2. 17. Several forms of DGS cells; a) spiral head; b) “H” shape; c) a square open-loop with a slot in middle section; d) open-loop dumbbell; e) interdigital DGS...	23
Figure 2. 18. The schematic of the DGS cell from (Y. Huang, 2013); “the orange indicates the copper and the white indicates the substrate’s insulation material”	24
Figure 2. 19. Microstrip line loaded by DGS cell; a) 3-D structure; b) microstrip transmission line; c) DGS cell etched in the ground plane; d) <i>S</i> -parameters response; f) the real part of the permeability; “the yellow and orange indicates the copper and the white indicates the substrate’s insulation material”	25
Figure 2. 20. Picture of the fabricated 1:6 Wilkinson power divider with DGS designed by (Oh et al., 2007)	26
Figure 3. 1. Overall design process of the thesis.....	27
Figure 3. 2. Filter design process	28
Figure 3. 3. Conventional design process of the DGS cells.....	29
Figure 4. 1. SIW transmission line; a) upper plane; b) bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”	31
Figure 4. 2. The simulated <i>S</i> -parameters using CST.....	31
Figure 4. 3. DGS cell modeling using microstrip TL; a) Cell schematic; b) 3D-Structure; “the yellow and orange indicates the copper and the gray indicates the substrate’s insulation material”	33
Figure 4. 4. DGS cell simulated <i>S</i> -parameters using CST	34
Figure 4. 5. 1 st order SIW-DGS BPF; a) upper plane; b) bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”	35
Figure 4. 6. <i>S</i> -parameters corresponding to the 1 st -Order SIW-DGS BPF	36
Figure 4. 7. 3 rd order SIW-DGS BPF; a) upper plane; b) bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”	37
Figure 4. 8. <i>S</i> -parameters "simulation results" of the 3 rd order SIW-DGS BPF	38
Figure 4. 9. 3 rd order SIW-DGS BPF with one dumbbell in the bottom plane; a) upper plane; b) bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”	39

Figure 4. 10. 3 rd order SIW-DGS BPF - with one Dumbbell cell simulated S-parameters	40
Figure 4. 11. 3 rd order SIW-DGS BPF - with three dumbbells; a) Upper plane; b) Bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”	41
Figure 4. 12. Simulation results of the 3 rd order SIW-DGS BPF with three dumbbells; a) S-parameters; b) Group-delay	42
Figure 4. 13. 3 rd order SIW-DGS C-Band BPF; a) Upper plane; b) Bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”	45
Figure 4. 14. 3 rd order SIW-DGS C-Band BPF; a) S-parameters; b) Group-delay	46
Figure 4. 15. 5 th order filter in C-band realized by using single-mode TM resonators (top) and relevant full-wave simulated response simulate with HFSS (bottom) (Tomassoni et al., 2017).....	47
Figure 5. 1. Fabricated X-Band filter; a) Upper plane; b) Bottom plane.....	48
Figure 5. 2. SMA connector and its corresponding dimensions	49
Figure 5. 3. Bad installation of SMA connector; a) upper face; b) bottom face.....	50
Figure 5. 4. Measured S_{21} response.....	51
Figure 5. 5. Ideal simulation results vs. measurement results.....	52
Figure 5. 6. SMA inclusion in the simulation; a) upper view; b) bottom view; c) side view	53
Figure 5. 7. Simulated S-parameters taking into account SMA bad installation vs. measured S-parameters	53
Figure 5. 8. . Field distribution due to SMA miss-installation; a) E-field top view; b) E-field bottom view; c) E-field side view; d) H-field top view; e) H-field bottom view; f) H-field side view;	56
Figure 5. 9. Fixing/Correcting SMA connector installation	57
Figure 5. 10. Measured Results after the correction of SMA installation; a) S_{21} measured response; b) S_{11} measured response.....	58
Figure 5. 11. Ideal simulation vs. measured S-parameters.....	58
Figure 5. 12. Inclusion SMA connectors in the Simulator; a) top view; b) oblique view of the upper face; c) oblique view of the bottom face.....	60
Figure 5. 13. Simulation SMA- included vs Measured results	60

Figure 5. 14. High-performance end launch connectors; **a)** high-performance SMA connector serve up to 26.5 GHz; **b)** 2.92mm or k- connector serves up to 50 GHz..... 62



LIST OF TABLES

Table 2. 1. Electromagnetic equations of the the TE modes in rectangular waveguide in phasor form	10
Table 2. 2. Characteristics comparison between waveguides, planar transmission lines, and SIW.....	15
Table 4. 1. Dimensions of X-band SIW & Taper which illustrated in Figure 4.1	31
Table 4. 2. DGS cell parameters.....	34
Table 4. 3. Center Dumbbell DGS dimensions	40
Table 4. 4. Port Dumbbells dimensions.	42
Table 4. 5. C-Band filter dimensions	45

1. INTRODUCTION

1.1. Introduction

Filters are passive components and considered the fundamental component of any communication system. Filters are a two-port frequency-selective networks that allow desired signal to go through (passband) while block the remaining undesired frequency (stopband). They have a wide application in radar and communication systems. Commonly used in transceiver circuits, as shown in Figure 1.1.

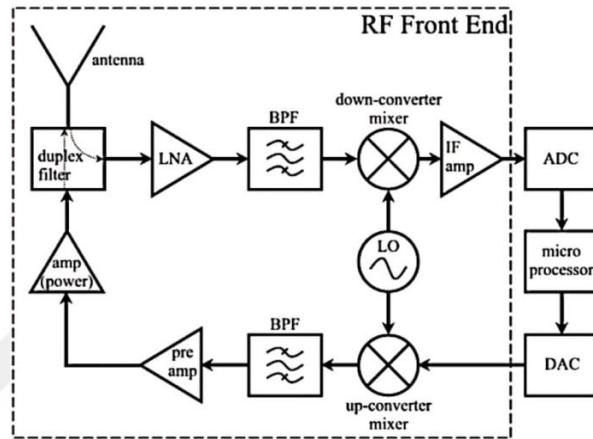


Figure 1. 1. Simplified RF front end transceiver block diagram

Filters are classified in terms of frequency selection into four categories, low-pass, high-pass, band-pass and band-stop, as shown in Figure 1.2. Also, it may be classified according to response (Chebyshev, Maximally flat, Elliptic, ...etc.) or technology (Lumped, Waveguide, Microstrip, ...etc.) or according to the frequency band (Narrowband or Broadband).

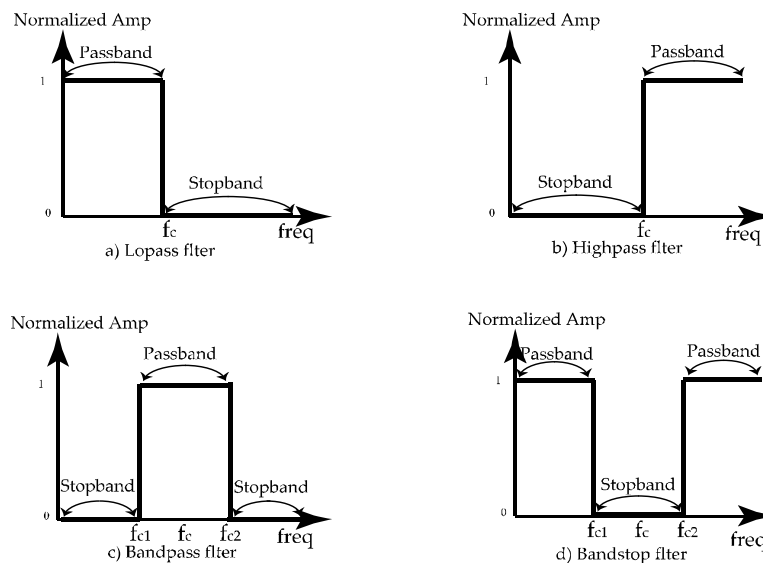


Figure 1. 2. Filter types according to frequency selectivity

The ever-increasing development and appearance of the new applications such as wireless communications keep challenging RF/microwave filters with ever more strict requirements; better performance, miniaturized, lighter weight, integrable and lower cost.

The lumped circuit element technology valid for the lower segments of the radio frequency ranging from 300 MHz to 3 GHz (HONG, 2011), lumped circuit element approximations of circuit theory valid at low RF frequencies, that is because at very high frequency there is high energy loss of the lumped elements and a severe radiation of the transmission lines (Laboratory, 1964). The distributed elements concept is applied for high RF and microwave frequencies (3 – 300) GHz, where the phase of the voltage or the current changes significantly over the physical extent of the device, where device dimensions are comparable to the electrical wavelength. As a result, the voltages and currents behave as wave inside the device, and hence the coupling effects affect the design process of components (HONG, 2011).

Passive Band-Pass Filters are a selective component which allows certain frequency band to pass through while, block the spectrum located above and below the desired frequency band. Nowadays, accelerated improvement in computer-aided design (CAD) tools, such as full-wave electromagnetic (EM) simulators concurrent with the precise fabrication technologies and developed measurement tools, have fundamentally enhanced the filter design. Many novel microstrip filters with advanced filtering characteristics have been demonstrated. Substrate integrated waveguide (SIW), early introduced by Ke Wu et al., has attracted plenty of attention and utilized in different applications due to its merits of higher quality factor, easy integration, low cost, high power handling and mass production (D. Deslandes & K.Wu, 2006; Y. Cassivi, 2002).

1.2. Problem Statement

In the modern microwave and millimeter-wave communication systems, highly integrated wideband bandpass filters play important roles for frequency selection. Meanwhile, the area of research focus on low loss, easily integrable and miniaturized and low-cost filters. General requirements that must be fulfilled in order match the increased demand on high-performance filters are low insertion loss, out of band rejection, high power capability, ease and cost of fabrication, and easy integration with other components. The commonly used technologies are rectangular waveguides and the microstrip technology. As it known non-planar metallic waveguides represented mainly by rectangular waveguide (RWG) shown in Figure 1.3. provides high-performance requirements in term of low loss high quality factor, however; its bulky, require precise fabrication processes and it is costly expensive in addition to integration difficulties.

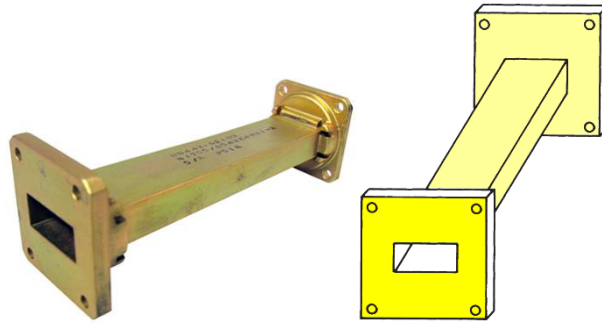


Figure 1. 3. Rectangular waveguide samples

Regarding planar technologies such as microstrip and coplanar waveguides shown in Figure 1.4. It distinguished by its lightweight and compactness moreover ease of fabrication and integration and cost-effectiveness. However; there are a lot of cons such as high losses, conductor losses due to high current densities in the tiny transmission space and high dielectric loss, accompanied by high electromagnetic (EM) radiation loss cases electromagnetic (EM) interference (high cross-talk) due to its unbounded structure, the low power handling, and low Q-factor.

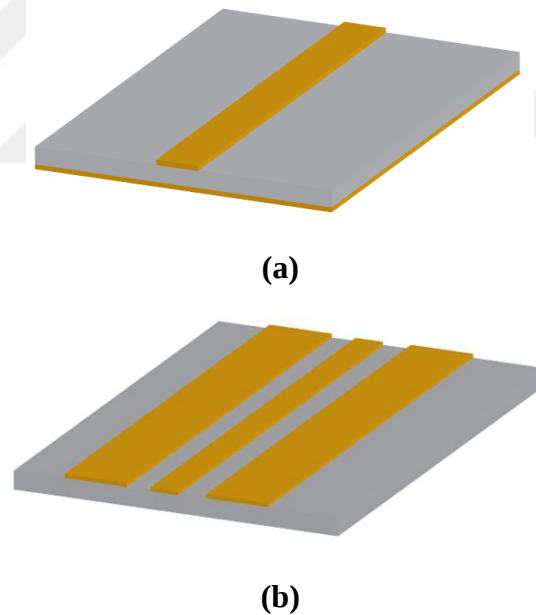


Figure 1. 4. Planer waveguides; **a)** Microstrip; **b)** coplanar waveguide

Traditional technology cannot satisfy all of the required aspects at once. There is a performance gap between traditional technologies. What about the solution that comes to fill the gap between traditional technologies? Recently, a hybrid guiding structure synthesizing the waveguide structures into a planer structure called substrate integrated waveguide (SIW) illustrated in Figure 1.5. First presented by (Hirokawa & Ando, 1998), and introduced by (Deslandes & Wu, 2001). SIW Provides the best compromise that can

fill the gap between the planar and non-planar waveguiding technologies, where it exploiting advantages adapted from both rectangular waveguides RWGs and microstrip lines. Also, it is featured by the ease of fabrication on a low loss profile dielectric substrate using regular PCB's fabrication processes.

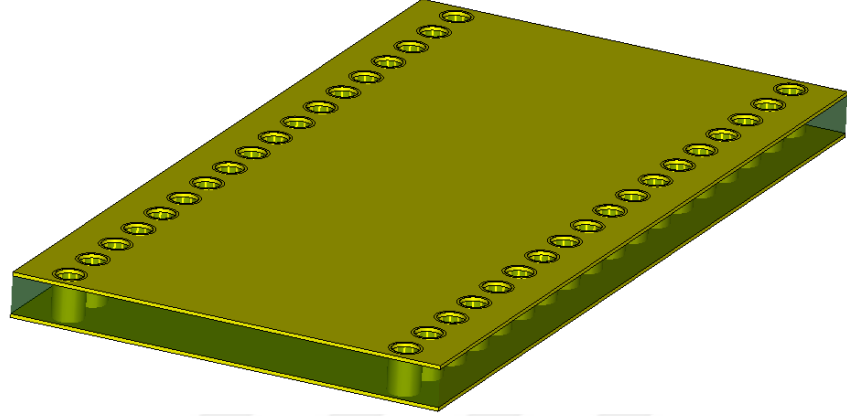


Figure 1. 5. Substrate integrated waveguide's general structure

1.3. Objectives

In this thesis, we focus on design two SIW filters, works on X and C bands. We illustrate our general procedure by designing two filters one for X-band general applications and the other proposed for C-band space communication. An initial design is developed by applying empirical formulas and design criteria available in the literature for SIW. A full-wave electromagnetic simulation of the initial design is realized. We optimize the SIW model using direct EM full wave simulation, which is implemented in two stages: optimizing the SIW width to achieve the desired low cutoff frequency, followed by the transition optimization to minimize reflections of the operating bandwidth. Then we investigate the effects of the Defected ground structures (DGS) cells by etching the cells under a microstrip transmission line to perform fast simulation, and to obtain the initial dimensions of the cells that provide the desired bandgap effect, and then we can use these cells as initial cells to be used with the SIW structure to construct the bandpass filter. Recently, a significant number of bandpass filters that realized by using the SIW and DGS architecture appeared, some examples of these filters are mentioned in (L. H. Weng, 2008; Y. Huang, 2013).

In SIW-DGS filters designed by (Liu, 2017; Z. He, 2013), the lower edge frequency of the proposed BPF filter controlled by SIW's width while the upper cutoff frequency is mainly controlled by the DGS cells. We illustrate our general procedure by designing two filters one for X-band general applications and the other proposed for C-band space communication.

1.4. Scopes

- Design and analyze the results by using EM-full wave simulator “CST Studio Suite 3D EM simulation and analysis software”, by analyzing the response referring to scattering parameters and group delay.
- Analyze the result by comparing simulation and measurement for other filters and proposed filters.
- The X-band filter is centered at 10 GHz with a fractional bandwidth FBW=40% while the C-band has a central frequency of 6.175 GHz with FBW=8.3%.
- The SIW-DGS proposed filters are built using the Diclad 880 laminate of a thickness of 0.508 mm, dielectric constant ϵ_r 2.2 and loss tangent 0.0009 fabricated using LPKF s104 milling machine, and the silver-plated vias using plated through-hole technique.

1.5. Organization of the thesis

The outline of this thesis is as follows:

Chapter 1 presents the introduction part of the thesis such as glance about filters in the communication systems, the problem statement, objectives of the thesis, the scopes of the thesis.

Chapter 2 provides the fundamental theories that cover the background of the work presented in the subsequent chapters. In which it starts with the general background theory of microstrip moving to rectangular waveguides. After that touches on the theory of SIW and finally the design defected ground structures (DGS).

Chapter 3 is related to the methodology included in block diagrams of the design procedure followed in this thesis.

Chapter 4 looks into filters design procedures in details.

Chapter 5 presents the fabricated filters with the measured results, also contains a comparison between the measured and simulated results to show their agreements. Finally, Chapter 6, Sums up with conclusion and recommendations.

2. THEORETICAL BACKGROUND

Through this Chapter, a background of basic transmission lines theory related to the waveguiding structures and transmission line types used in filters design will be presented in addition to the defected ground structure (DGS) theory.

The first section deals with the microstrip transmission lines, moving to the rectangular waveguide and the related theory, and then the substrate integrated waveguides (SIW).

In the second section, we discuss the transitions between the microstrip and SIW. In the third section we touch the theory of defected ground structures (DGS).

2.1. Transmission Lines

2.1.1. Introduction

The transmission line (TL) circuit is simply a structure capable to carry and guide electrical signals from one point to another. The propagating EM waves store the energy in both electric and magnetic fields, while the transmission lines stores electric and magnetic energy in the form of distributed capacitance and inductance (Steer, 2016).

As mentioned earlier, the main category of transmission lines divides it into planar and non-planar lines. Planar transmission lines include microstrip lines, coplanar waveguides, slot lines, and strip lines. Whereas examples of non-planar transmission lines include waveguides and coaxial lines.

Recently, the intensive research on SIW led to present it as a new type of waveguiding technologies, that enhanced the horizons of development level of circuits design associated with microwave and millimeter-wave applications and allowed to apply the non-planar circuits concepts into planar structures. Moreover, SIW is considered as a planar structure easy to integrate with other planar structures, with performance comparable to waveguide performance.

2.1.2. Microstrip lines (HONG, 2011)

Microstrip structure is the most popular planar type transmission line formed by conducting strip having a width (W) and a cladding thickness (t) formed over the upper side of the thin dielectric substrate with a relative permittivity (ϵ_r) and substrate thickness (h), while, the ground is the bottom conducting layer as shown on Figure 2.1.

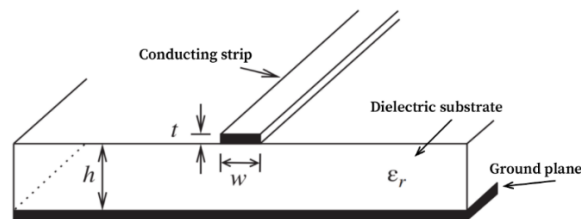


Figure 2. 1. Microstrip line structure

Microstrip supports the Quasi-TEM mode as in Figure 2.2, due to its non-homogeneous construction, that the field between the transmission strip and the ground, exists within two different media, in the air (the fringing field) and the dielectric medium. As for the TEM wave, it has only transverse components and the propagation velocity of the TEM waves depends on the properties of the material including, the permittivity ϵ and the permeability μ .

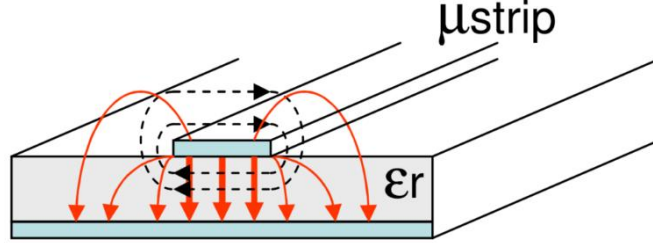


Figure 2. 2. Microstrip quasi-TEM mode where the dashed lines represent the magnetic field lines and the solid lines represent the electric field lines

As mentioned before in microstrip lines the operating mode is the quasi-TEM mode which appears because there are components of the electric and magnetic field propagates through the air above the top pattern and through the dielectric substrate. Due to this, two different mediums having different characteristics which results in wave propagation with different speeds in both the regions. This is referred as quasi TEM mode.

Due to field existence into two different media, the effective dielectric constant ϵ_{re} is used, and characteristic impedance Z_c for microstrip lines with very thin conductors derived by (Hammerstard, 1975) provided in a closed-form presented in the Equations (2.10 – 2.13) :

For $W/h \leq 1$:

$$\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left\{ \left(1 + \frac{12h}{W} \right)^{-0.5} + 0.04 \left(1 - \frac{W}{h} \right)^2 \right\} \quad (2.10)$$

$$Z_c = \frac{\eta}{2\pi\sqrt{\epsilon_{re}}} \ln \left(\frac{8h}{W} + \frac{W}{4h} \right) \quad (2.11)$$

where η is the wave impedance in free space and equal to $120\pi \Omega$.

For $W/h \geq 1$:

$$\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + \frac{12h}{W} \right)^{-0.5} \quad (2.12)$$

$$Z_c = \frac{\eta}{\sqrt{\epsilon_{re}}} \left\{ \frac{h}{W} + 1.393 + 0.677 \ln \left(\frac{W}{h} + 1.444 \right) \right\}^{-1} \quad (2.13)$$

Where ε_r , h and W represent the dielectric constant of the substrate, the height of the substrate, and the width of the microstrip line respectively.

The guided wavelength of the quasi-TEM mode of microstrip is given by Equation 2.14 and Equation 2.15

$$\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_{re}}} \quad (2.14)$$

$$\lambda_0 = \frac{c}{f} \quad (2.15)$$

c is the velocity of light in free space equal to $c \approx 3.0 \times 10^8$ m/s, λ_0 is the free-space wavelength corresponding to the operating frequency and the electrical length θ is given by Equation (2.16)

$$\theta = \beta \cdot l \quad (2.16)$$

where β is the propagation constant $\beta = 2\pi/\lambda_g$, and l represent the physical length of the Microstrip line.

Microstrip line components have the advantage of the ease of design, fabrication, integration in addition to the compact size. There is no cutoff frequency for the TEM mode. Microstrip can transmit signals from direct current (DC) up to high frequency. the aforementioned features led to make it widely used technology in RF and microwave circuit design. On the other hand, microstrip suffers from high losses and low power handling capacity. Also, susceptible to cross-talk and unintentional radiation due to microstrip's unbounded structure.

Microstrip circuits as a final product are supplied in a metallic enclosure work as Shield against radiation and electromagnetic interference, this shield affects the effective dielectric constant and hence the characteristic impedance. The rule of thumb to reduce the enclosure effect in filter design suggested by (LANCASTER, 2001) which state that the height of the top conducting shield should be more than eight times the substrate thickness and the distance to sidewalls of the shield at least five times more than the substrate thickness.

2.1.3. Rectangular Waveguides

Mots of transmission lines are formed with two-conductor and they support *TEM* mode, where the electric and magnetic fields are perpendicular to the propagation direction at any point, also they can transmit signals from dc up to high frequencies. The metallic waveguides are typically single enclosed conductors filled with an insulating medium as shown in Figure 2.3. Most of the Waveguides are air-filled type, it has better performance than other dielectric-filled waveguides, other types filled by materials of dielectric constants of a value higher than air dielectric constant are used for size reduction purposes, which mean the Waveguide's size decreases as the relative dielectric constant of the filling dielectric increases. RWG does not support TEM mode because it needs more than a single conductor to exist, waveguide supports transverse electric (*TE*) or transverse magnetic (*TM*) modes, or a combination from them called hybrid mode. They can propagate the signals at frequencies above the dominant *TE* or *TM* mode cut off frequency (Pozar, 2012). In which, at a given operating frequency f , only those modes with $f > f_c$ can propagate, as shown in Figure 2.4 f_c represents the cutoff frequency, is the frequency at which attenuation occurs below and propagation takes place at greater frequencies (High Pass).

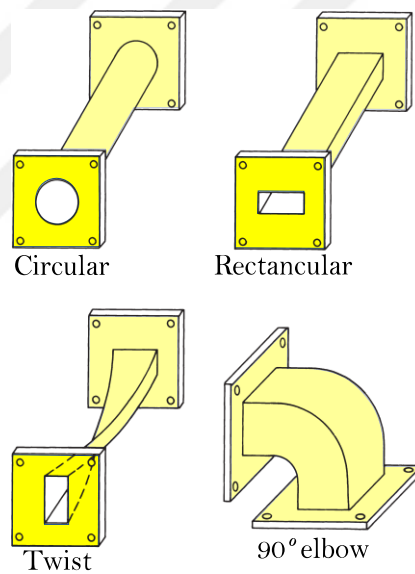


Figure 2. 3. Typical waveguides; **a)** circular waveguide; **b)** rectangular waveguide; **c)** twist; **90-degree** elbow

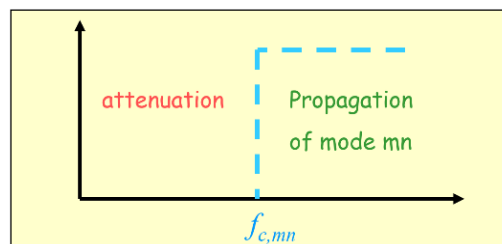


Figure 2. 4. Waveguide cutoff frequency

Waveguides utilized at high frequencies because they provide lower signal attenuation and larger bandwidth than transmission lines. Also, it is selected for an application that requires high Q-factor moreover the ability to handle high power levels since the fields of propagating waves are spread more uniformly over a larger cross-sectional area than tiny cross-section transmission lines. Nevertheless, waveguides are impractical due to large size and high manufacturing costs, also they suffer from integration problems with planer circuits.

The most used waveguides are rectangular cross-sections waveguides, typical structure is shown in Figure 2.5 where a and b are the inner dimensions.

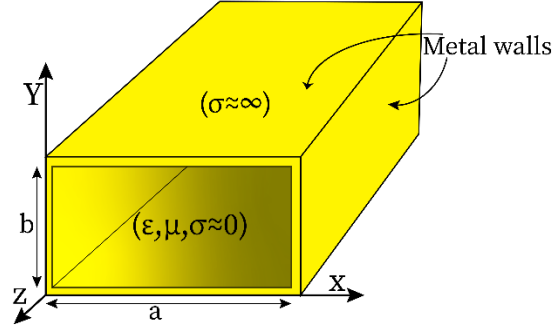


Figure 5. Geometry of a rectangular waveguide with perfectly conducting walls, filled with a lossless material

Considering the propagation is in Z-direction, TE modes ($E_z = 0, H_z \neq 0$) stands for transverse electric, in which the electric field is transverse to the direction of propagation, and TM modes ($E_z \neq 0, H_z = 0$), transverse magnetic, where the magnetic field is transverse to the direction of propagation. As for hybrid modes ($E_z \neq 0, H_z \neq 0$), HE modes all component exist in the direction of propagation.

The Electromagnetic equations of the electric and magnetic fields in each mode are derived and stated in (Sadiku, 2018). The equations (2.17-2.28) related to TE modes inside waveguide are summarized in Table 2.1.

Table 2. 1. Electromagnetic equations of the TE modes in rectangular waveguide in phasor form

TE modes governed equations	
$E_{xs} = -\frac{j\omega\mu}{h^2} \left(\frac{n\pi}{b}\right) H_0 \cos\left(\frac{m\pi x}{a}\right) \sin\left(\frac{n\pi y}{b}\right) e^{-\gamma z}$	2.17
$E_{ys} = -\frac{j\omega\mu}{h^2} \left(\frac{m\pi}{a}\right) H_0 \sin\left(\frac{m\pi x}{a}\right) \cos\left(\frac{n\pi y}{b}\right) e^{-\gamma z}$	2.18
$E_{zs} = 0$	2.19

$$H_{xs} = \frac{j\beta}{h^2} \left(\frac{m\pi}{a} \right) H_0 \sin \left(\frac{m\pi x}{a} \right) \cos \left(\frac{n\pi y}{b} \right) e^{-\gamma z} \quad 2.20$$

$$H_{ys} = \frac{j\beta}{h^2} \left(\frac{n\pi}{b} \right) H_0 \cos \left(\frac{m\pi x}{a} \right) \sin \left(\frac{n\pi y}{b} \right) e^{-\gamma z} \quad 2.21$$

$$H_{zs} = H_0 \cos \left(\frac{m\pi x}{a} \right) \sin \left(\frac{n\pi y}{b} \right) e^{-\gamma z} \quad 2.22$$

$$\Upsilon = \dot{u} \sqrt{\left(\frac{m\pi}{a} \right)^2 + \left(\frac{n\pi}{b} \right)^2 - k^2}, \quad k = \omega \sqrt{\mu \epsilon} \quad 2.23$$

$$\eta = \frac{\sqrt{\frac{\mu}{\epsilon}}}{\sqrt{1 - \left(\frac{f_c}{f} \right)^2}} \quad 2.24$$

$$f_c = \frac{\dot{u}}{2} \sqrt{\left(\frac{m}{a} \right)^2 + \left(\frac{n}{b} \right)^2} \quad 2.25$$

$$\lambda_c = \frac{\dot{u}}{f_c} \quad 2.26$$

$$\beta = \frac{\omega}{\dot{u}} \sqrt{1 - \left(\frac{f_c}{f} \right)^2} \quad 2.27$$

$$h^2 = \left(\frac{m\pi}{a} \right)^2 + \left(\frac{n\pi}{b} \right)^2, \quad \dot{u} = \frac{1}{\sqrt{\mu \epsilon}} \quad 2.28$$

In Table 2.1 the terms Υ represents the propagation constant, η represents the intrinsic impedance, f_c and λ_c are the cutoff frequency and the cutoff wavelength for the TE_{mn} mode, β represents the propagation phase constant of uniform plane wave, k is the wavenumber, and $\dot{u}$ represents the phase velocity.

It should be noted that the phase constant β and the wavenumber k are indeed equal in TEM modes but in other modes, they can be different.

In (TE_{mn}, TM_{mn}) modes each set of integers m and n gives a different field pattern or mode, referred to as TM_{mn} mode, in the waveguide. Integer m equals the number of half-cycle variations in the x -direction, and integer n is the number of half-cycle variations in the y -direction.

It is standard practice to have $a > b$, where TE_{10} mode is the dominant mode with the lowest cutoff frequency. TE_{10} mode has these properties: (1) the electric field is confined to planes perpendicular to the direction of propagation, whereas the magnetic field has both transverse and longitudinal components, and (2) the electric field is directed from one of the guide surfaces straight across to the opposite surface, as shown in Figure 2.6.a. Figures 2.6.b, 2.6.c demonstrate the top view of a simulated distribution of magnetic and electric fields of TE_{10} mode in a rectangular waveguide.

From Equations (2.17-2.22), we can obtain the field patterns for the TE modes. For the dominant TE_{10} mode, substituting $m = 1$ and $n = 0$ in that equations, we obtain the equation of TE_{10} which are the Equations (2.29-2.33).

$$H_{zs} = H_0 \cos\left(\frac{\pi x}{a}\right) e^{-j\beta z} \quad (2.29)$$

Changing from phasor to time domain,

$$H_z = \text{Re}(H_{zs}) \quad (2.30)$$

$$H_z = H_0 \cos\left(\frac{\pi x}{a}\right) \cos(\omega t - \beta z) \quad (2.31)$$

$$E_y = \frac{\omega \mu a}{\pi} H_0 \sin\left(\frac{\pi x}{a}\right) \sin(\omega t - \beta z) \quad (2.32)$$

$$H_x = -\frac{\beta a}{\pi} H_0 \sin\left(\frac{\pi x}{a}\right) \sin(\omega t - \beta z) \quad (2.33)$$

$$E_z = E_x = H_y = 0 \quad (2.34)$$

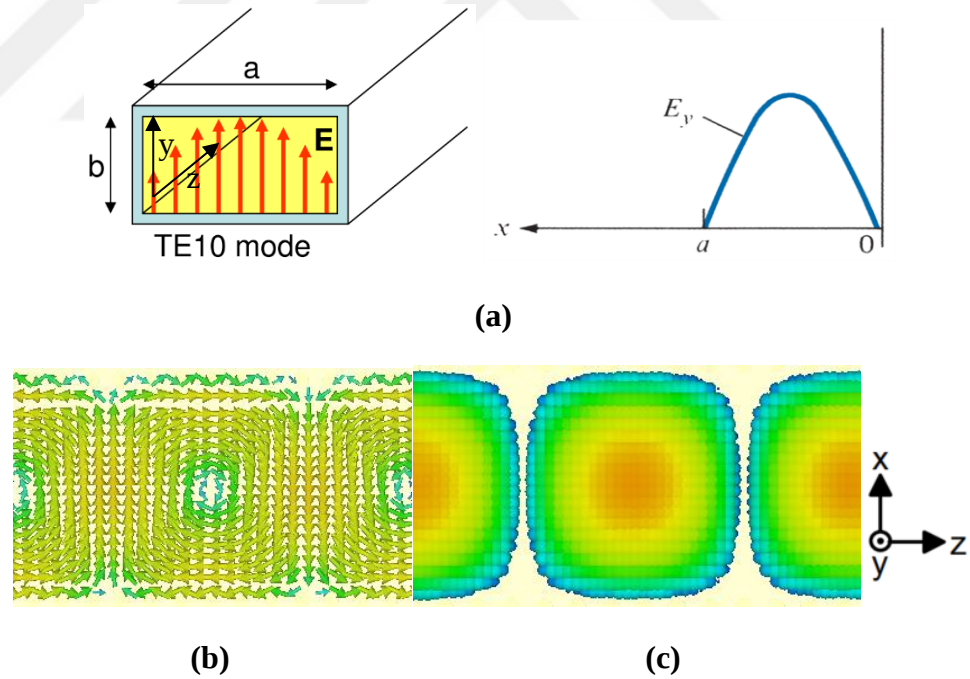


Figure 2. 6. Rectangular waveguide TE_{10} mode field distribution; a) Electric field lines (front view); b) Magnetic field (top view); c) Electric field (top view)

If more than one mode is propagating, the waveguide is overmoded. Single mode propagation is highly desirable to reduce dispersion. This occurs between cutoff frequency for dominant mode and cutoff frequency of next higher mode.

2.1.4. Substrate integrated waveguides

Substrate integrated waveguide (SIW) is one of emerging guided-wave structures. This novel waveguiding technology has become popular in the past decade, SIW principle originated from the dielectric-filled rectangular waveguide.

The necessity behind the development of SIW is the gap between traditional technologies, in which the Air-filled rectangular waveguides offers high power capacity and high Q-factor, but they are bulky and expensive, moreover, requires complex transitions to integrated planar circuits. On the other hand, the planar transmission lines are low profile (low profile means small height and width) structures with ease of integration with other planer circuits, but they lack the high-performance requirements with high losses and radiation leakage. SIW is a kind of integration a metallic rectangular waveguide into a planar microstrip substrate, can be constructed using standard planar fabrication processing techniques that include printed circuit board (PCB) for single-layer structures or other processing techniques for multilayer structures such as low-temperature cofired ceramic (LTCC) and liquid crystal polymer (LCP).

After looking at Figure 2.7 we can say that SIW is a straightforward integration of a rectangular waveguide into a microstrip substrate. This provides the best compromise between the microstrip and waveguide technology with reduced Q factor of the Air-filled waveguide due to dielectric filling the structure as well as the volume reduction, but the entire circuit including planar circuit, transition, and waveguide are constructed over single laminate (Deslandes & Wu, 2001). This recently motivated the scholars to develop SIW as we can notice from (Ke, Deslandes, & Cassivi, 2003). A Well fabricated SIW can work up to 150 GHz which enables it to serve Terahertz Applications (Wang, Hao, & Kui-Kui, 2016), also can work up to higher frequencies depending on fabrication ability.

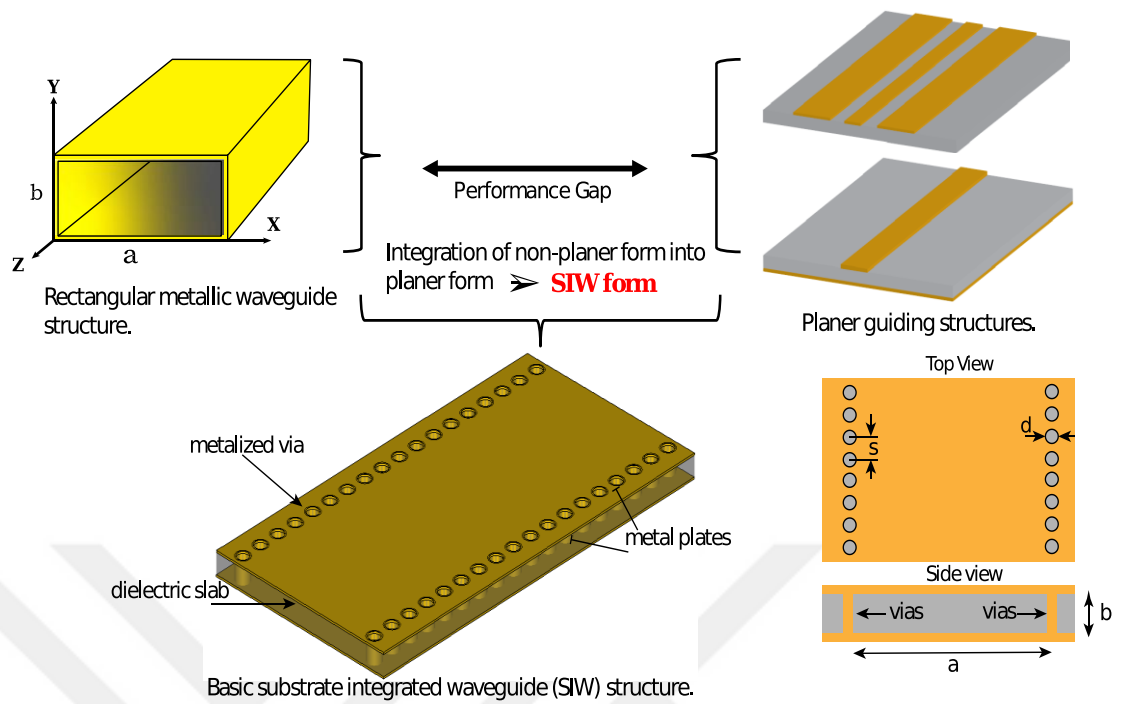


Figure 2. 7. Performance gap between planar and non-planar structures led to SIW development

SIW is a synthesized version of a dielectric-filled rectangular waveguide (RWG), embedded on low loss profile dielectric substrate through PCB, where the top and bottom are represented by the laminate's upper and lower cladding of the laminate, while the side walls are constructed using two-row (chains) of metalized vias (cylinders) that form electrical side walls prevent the field to pass through in the same principle as in the conventional rectangular waveguides, as described on Figure 2.8.

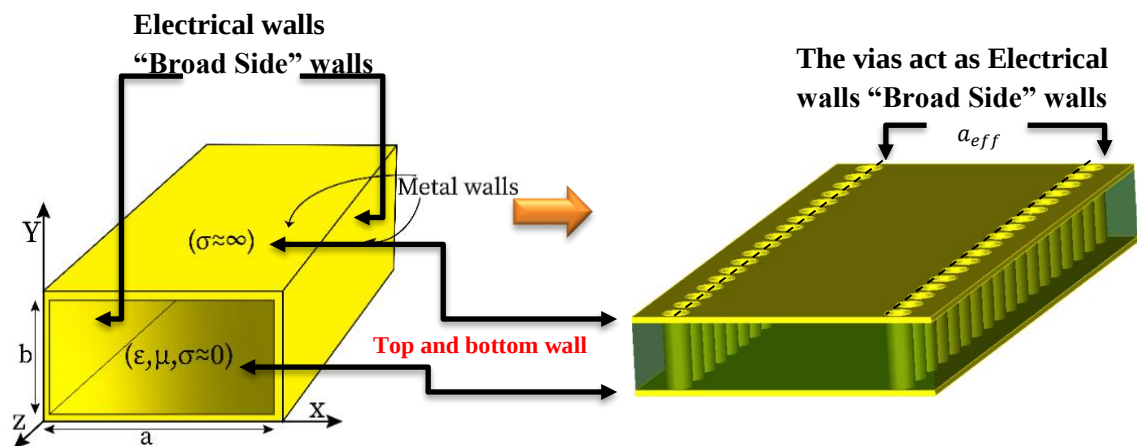


Figure 2. 8. SIW and its equivalent RWG

This new architecture has characteristics that mix the affirmative characteristics from the microstrip and the rectangular waveguide as illustrated in Table 2.2.

Table 2. 2. Characteristics comparison between waveguides, planar transmission lines, and SIW

	Cost	Loss	Power handling capability	Compactness	Self-shielded	Self-packaged
Waveguides	☹️	😊️	😊️	☹️	😊️	😊️
Planar Transmission lines	😊️	☹️	☹️	😊️	☹️	☹️
SIW	😊️	😐️	😐️	😊️	😊️	😊️

Principally, Substrate Integrated Waveguide (SIW) obey principles similar to the basic principles of conventional rectangular waveguides, with some differences. SIW side walls represented by truncated vias that do not allow longitudinal current flow (Chen & Wu, 2014), thus SIW does not support Transverse Magnetic (TM) modes as in a conventional metallic waveguide. Since most of the times, the SIW width is relatively greater than substrate thickness, the dominant mode of SIW is TE_{10} as it presented in Figure 2.9, followed by TE_{20} as the next higher mode.

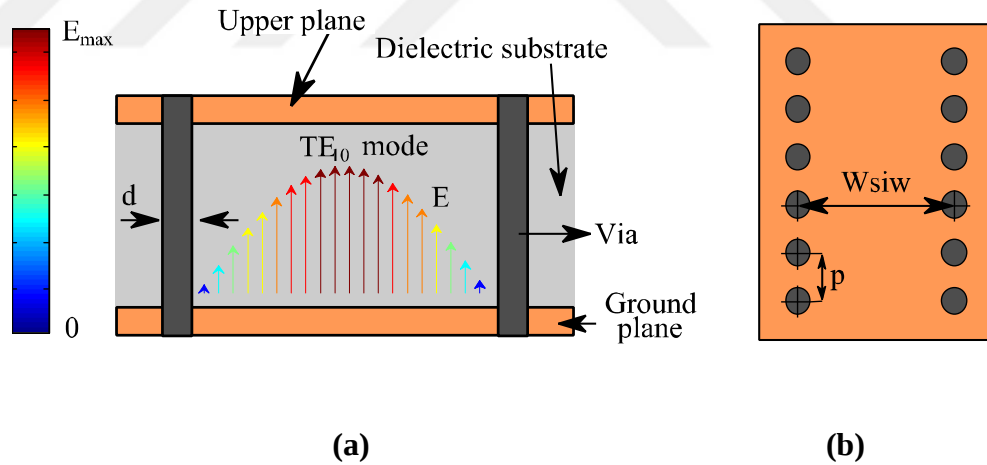


Figure 2. 9. SIW transmission line with TE_{10} mode; **a)** cross-section; **b)** top view

In the design of SIW, the determined parameters are d, P, a representing respectively the diameter of metalized via-holes, the center to center distance between adjacent via-holes, and the width of SIW, Shown in Figure 2.10. These parameters must be chosen carefully in order to avoid leakage losses and avoiding the bandgap “stopband” effects that appear in the periodic structures are subject to, moreover, these parameters must be chosen to maintain the structure robustness and avoiding overproportion, to

avoid all of these effects design rules stated by (Deslandes & Ke, 2006) must be respected;

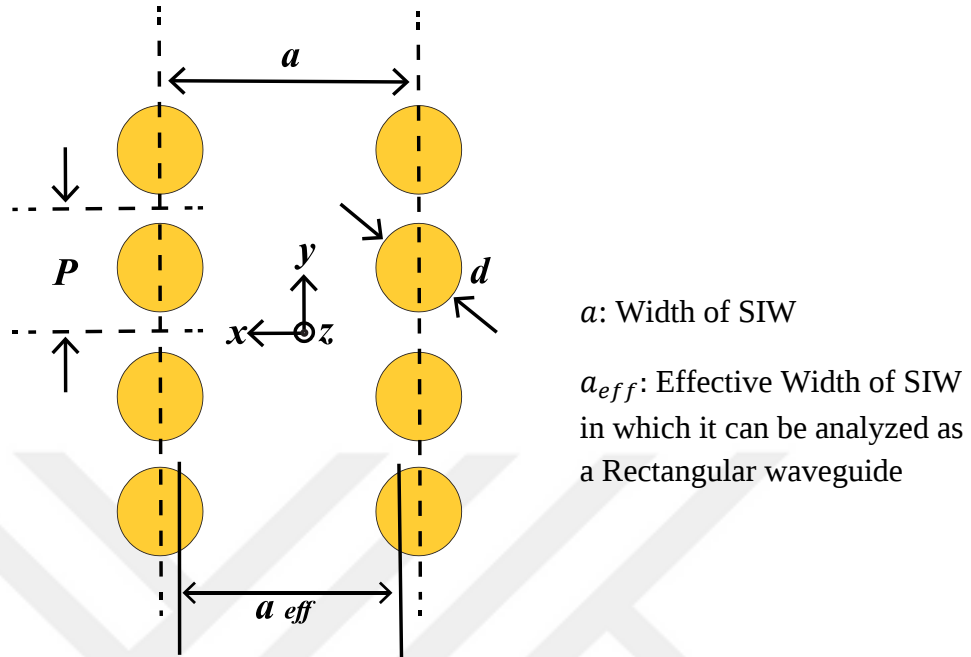


Figure 2. 10. Top view of the SIW

- To ensure mechanical toughness (to avoid Over perforation) of SIW Equation 2.35 must be satisfied.

$$P/\lambda_c > 0.05 \quad (2.35)$$

- Periodic structures are subject to electromagnetic bandgaps or stopband effects, to ensure that there is no bandgap over the waveguide bandwidth of interest the criteria stated in Equation 2.36 must be satisfied.

$$P/\lambda_c < 0.25 \quad (2.36)$$

- Judging from the periodic gaps, the SIW structures are subject to a possible field leakage problem, to avoid this problem the criteria stated in Equation 2.37 must be followed.

$$P \leq 2d \quad (2.37)$$

The effective width of SIW is the width that equivalent to the width of an equivalent Rectangular waveguide (RWG), in which both the SIW and RWG have the same dominant cutoff frequency. The SIW's effective width formula expressed in Equation 2.38 presented by (Y. Cassivi, 2002), using this effective width the SIW can be analyzed as a rectangular waveguide.

$$a_{eff} = a - \frac{d^2}{0.95 * p} \quad (2.38)$$

A more accurate empirical equation to calculate the SIW's effective width stated by (Feng & Ke, 2005) is presented in Equation 2.39.

$$a_{eff} = a - 1.08 \times \frac{d^2}{p} + \frac{d^2}{a} \quad (2.39)$$

In filter design, to avoid dispersion effects, the operating bandwidth of SIW should be confined in between the cutoff frequency of the dominant mode TE_{10} and cut off frequency of the next higher mode TE_{20} (Y. Cassivi, 2002). The cutoff frequency of the TE_{10} mode is presented in Equation 2.40 while the cutoff frequency of the TE_{20} mode is presented in Equation 2.41.

$$F_{c(TE_{10})} = \frac{C_0}{2\sqrt{\epsilon_r}} \left(a - \frac{d^2}{0.95 * b} \right)^{-1} \quad (2.40)$$

$$F_{c(TE_{20})} = \frac{C_0}{\sqrt{\epsilon_r}} \left(a - \frac{d^2}{1.1 * b} - \frac{d^3}{6.6 * b^2} \right)^{-1} \quad (2.41)$$

C_0 is the speed of light in free space. The thickness of SIW affects only the Q - factor.

Figure 2.11, illustrates the effect of changing the width of the SIW "a" on the cut off frequency of the TE₁₀ mode, as we can notice that increasing the width leads to a decrease in the 3 dB cut off frequency.

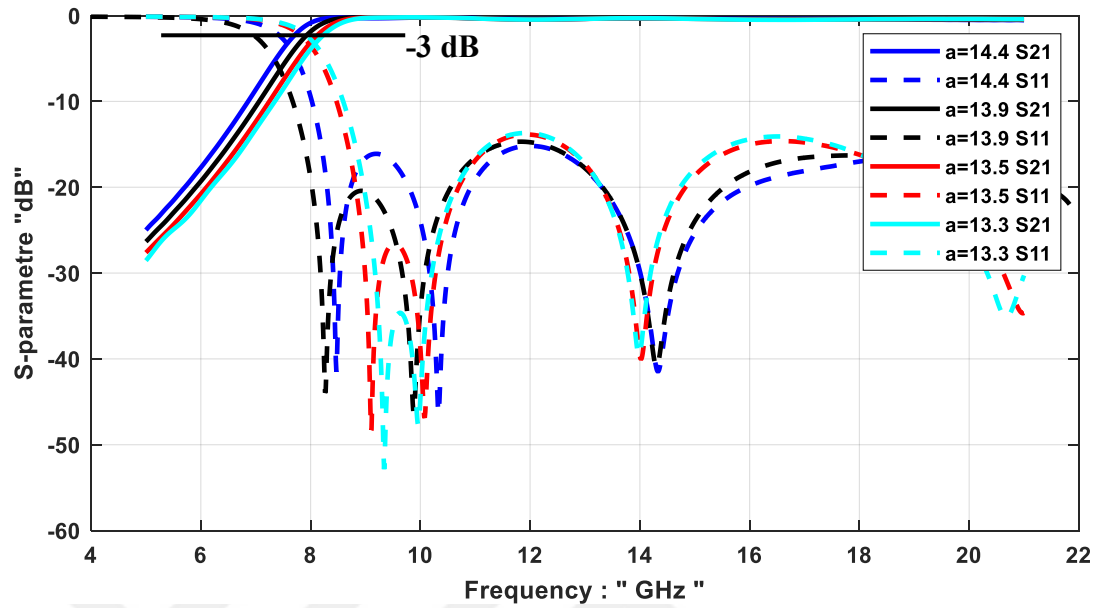


Figure 2. 11. Effect of changing the SIW's width (a) on the TE₁₀ mode cutoff frequency

2.2. Microstrip – SIW Transition

In order to connect and integrate SIW circuits either passive such as filters, couplers, power dividers, and antennas or active circuits such as amplifiers, mixers, oscillators with other planer circuits like microstrip or coplanar waveguide (CPW) Low-reflection transitions are required. Also, it is required to make the component design independent from the influence of the transition.

Microstrip-SIW transition is used for adapting the impedance and to excite SIW fundamental mode TE_{10} using the microstrip line *quasi* – TEM mode as shown in Figure 2.12.

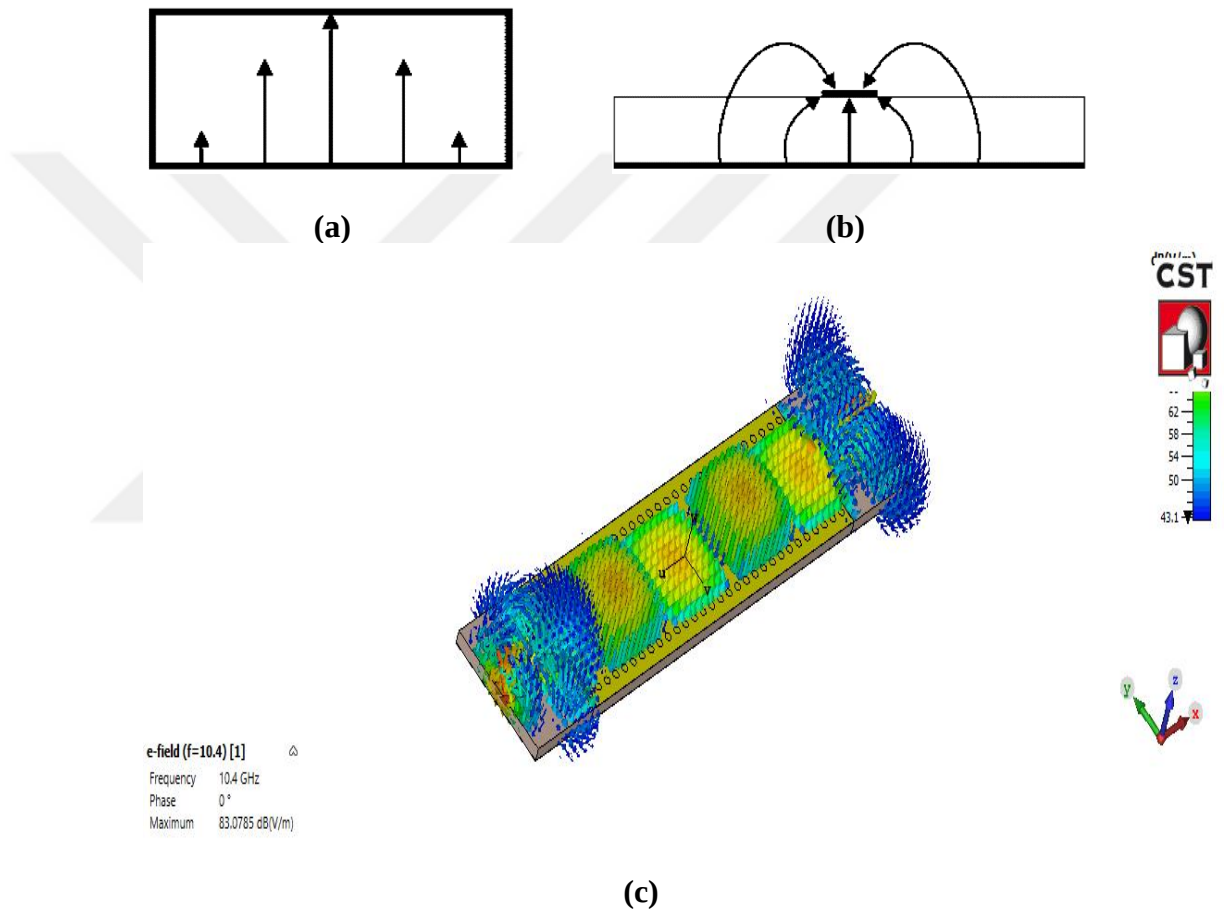


Figure 2. 12. SIW excitation using microstrip port; **a)** TE_{10} -SIW; **b)** *Quasi* – TEM microstrip line; **c)** Electric field distribution in SIW transmission line exploiting the TE_{10} mode excitation using Quasi-TEM mode

The first transition was the microstrip taper introduced by (Deslandes & Wu, 2001), and it still the most widely used type of Microstrip-SIW transition in single-layered circuits shown in Figure 2.13. Since it covers the complete bandwidth of the wideband SIW filters. Its performance is better when compared to other microstrip transitions such as direct transition, and easier to design than the Coplanar-SIW

transitions, moreover, well-optimized taper transition can provide minimum insertion and return losses. Most of the time, the taper transition physical dimensions are calculated using computational solutions available on Electromagnetic simulation programs due to the difficulty of yielding accurate analytical representation (Caleffo, 2016). The difference in the field modes between SIW (transverse electric field) and microstrip (transverse electromagnetic field) is considered as the main factor that increases the difficulty of calculation for accurate analytical solutions.

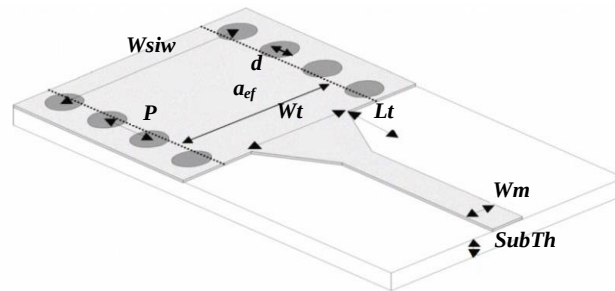


Figure 2. 13. Configuration of the microstrip-SIW transition

Microstrip Line - Substrate Integrated Waveguide new wideband transition introduced by (Kordiboroujeni & Bornemann, 2014) illustrated in Figure 2.14, this transition offer insertion and return losses better than the traditional transition in (Deslandes & Wu, 2001). The new aspect of this transition is the addition of two vias to the widely used microstrip taper transition.

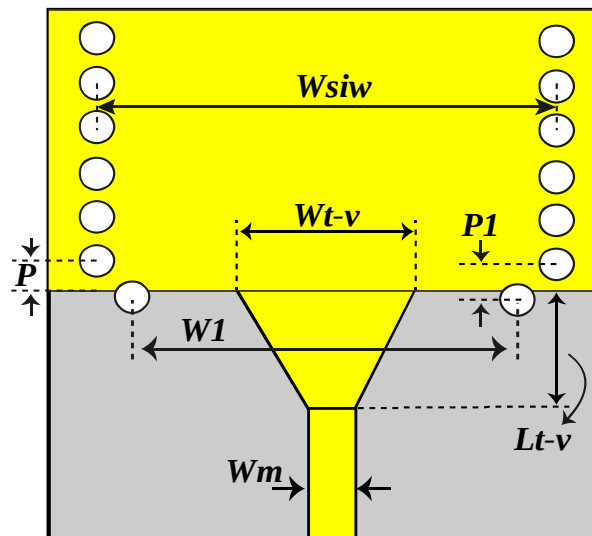


Figure 2. 14. Structural parameters of the new taper-via transition between microstrip and SIW (Kordiboroujeni & Bornemann, 2014)

After calculation of SIW and microstrip parameters for the desired frequency band, (Kordiboroujeni & Bornemann, 2014) extracted simple formulas to calculate the taper-via transition parameters based on their empirical experiments, the parameters are in Equations (2.42-2.45) as follow w_{t-v} , L_{t-v} , p_1 , w_1 , the width of the transition, the length of the transition, the distance between the last side vias and the transition vias, and distance separation between the transition vias as it depicted on Figure 2.14.

$$w_{t-v} = w_m + 0.1547 * W_{siw} \quad (2.42)$$

$$L_{t-v} = 0.2368 * \lambda_{g-ms} \quad (2.43)$$

$$p_1 = 0.6561 * p \quad (2.44)$$

$$w_1 = 0.8556 * W_{siw} \quad (2.45)$$

$$\lambda_{g-ms} = \frac{\lambda_{g0}}{\sqrt{\epsilon_{reff}}} \quad (2.46)$$

Where λ_{g-ms} in Equation 2.46 is the microstrip line guided wavelength calculated at the center frequency, λ_{g0} is the free space wavelength at the center frequency, and ϵ_{reff} is the effective dielectric constant of the microstrip line at the center frequency.

2.3. Defected Ground Structures

Defected ground structures (DGS) was first proposed by Korean scholar (Jong-Im et al., 1999) based on studying of Photonic-bandgap (PBG) structures. DGS is realized by etching of a simple structure “fault” in the ground plane, sometimes using more complicated shapes and it can be used in cascaded form for better performance. DGS cells were used first in the ground of transmission lines such as microstrip and coplanar lines. That's why it are called defected ground structures. DGS cells are disturbing the electrical current distribution which leads to changes in the transmission line characteristics such as equivalent capacitance and inductance (L. H. Weng, 2008). DGS presence usually provides a band-rejection feature (Ahn et al., 2001).

Some of popular simple DGS cells like square head slot, circular head slot, dumbbells slot, and arrowhead slot, are shown in Figure 2.15. DGS cells are utilized in microwave circuits such as antennas to enhance the bandwidth and increase the gain, also, in filters, power dividers, and couplers, for enhancing the performance and for shrinking elements size. DGS cells are being modeled by LC resonance circuits, in which the resonance frequency is controlled by the cell's shape and dimension.

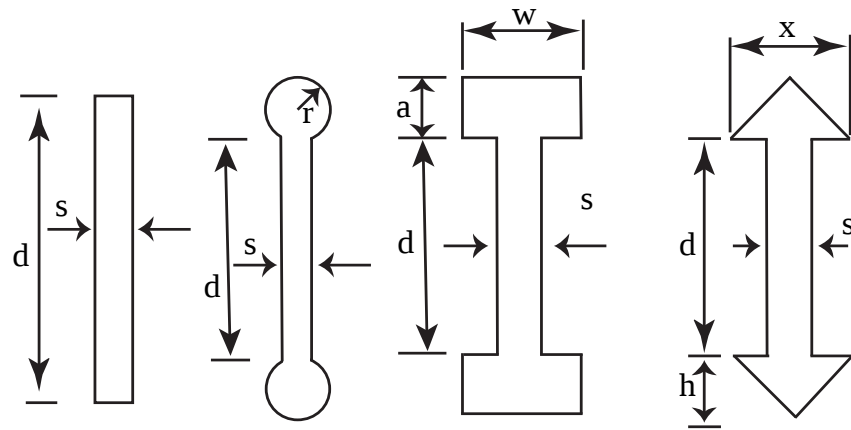
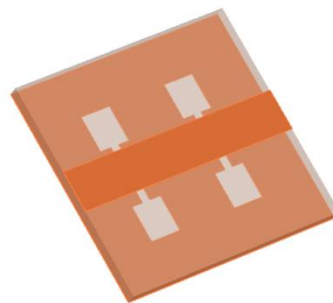


Figure 2. 15. Simple DGS slots

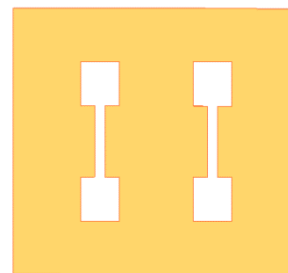
DGS cells can provide an electromagnetic bandgap, this feature is being used in microwave circuits in order to provide band stop, also used to provide low pass filter response, which offer efficient and compact geometry than using open-stubs or stepped line, simply by loading a microstrip line with DGS cells of appropriate characteristics in the ground plane of a microstrip line, we can get the low pass response as it is shown in Figure 2.16.



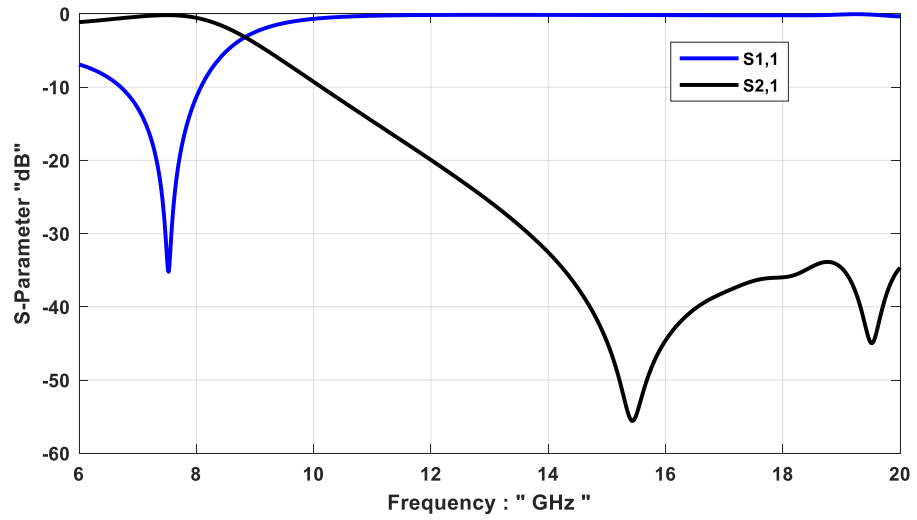
(a)



(b)



(c)



(d)

Figure 2. 16. Microstrip line loaded by dumbbell DGS cells (Jong-Im et al., 1999); **a)** LPF structure; **b)** microstrip transmission line; **c)** dumbbell DGS cells etched in the ground plane; **d)** S-parameters response

Different types of DGS cells have been used in filter design, mainly to enhance the stopband characteristics properties and suppress harmonics, and spurious response, also, for circuit miniaturizing purposes. A variety of relatively new configurations that were summarized by (L. H. Weng, 2008) are shown in Figure 2.17, these cells can provide better performance than those of simple shapes mentioned in Figure 2.15.

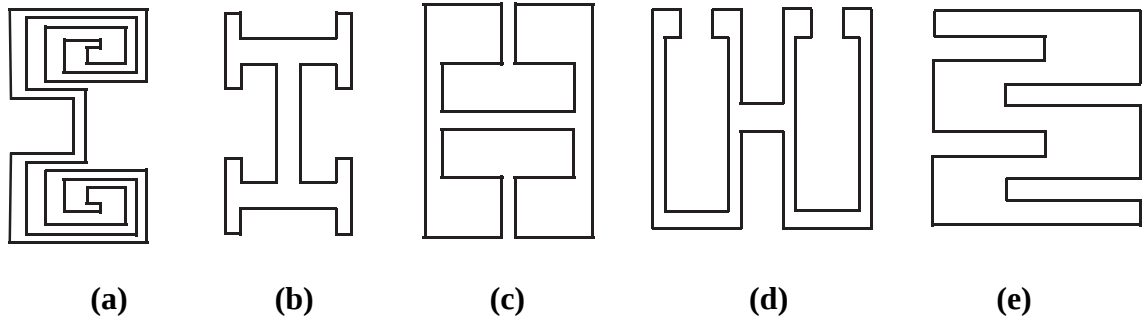


Figure 2. 17. Several forms of DGS cells; **a)** spiral head; **b)** “H” shape; **c)** a square open-loop with a slot in middle section; **d)** open-loop dumbbell; **e)** interdigital DGS

A novel DGS unit cell proposed by (Z. He, 2013) offers better performance compared with previous mentioned DGS cells, it has an advantage distinguish it from traditional cells, So that it can be incorporated in the top plane of SIW structures, which is more convenient with SIW structures than etching in the ground plane. The resonant frequency of the cell is controlled by the geometry and dimensions of the cell (Y. Huang,

2013). Figure 2.18. illustrate the schematic of the cell. It is reported that the bandgap effect may be due to DGS's negative permeability in the vicinity of its resonant frequencies.

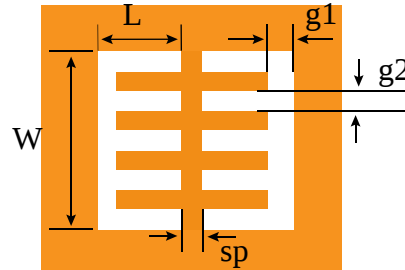
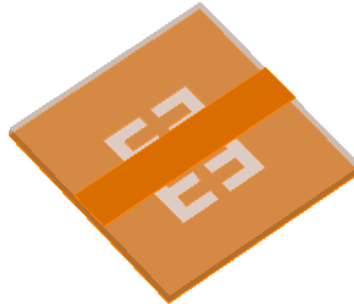


Figure 2. 18. The schematic of the DGS cell from (Y. Huang, 2013); “the orange indicates the copper and the white indicates the substrate’s insulation material”

The conventional method for analyzing the DGS which based on trial and error iterative method is used to analyze DGS cells to obtain the required dimensions of the cells that give the intended response using the full-wave simulator, to conduct a fast simulation, we incorporate the DGS cell with a microstrip transmission line, which reduce number of the cells in the simulator mish, after that we use these DGS cells with SIW structure. Figure 2.19 illustrates the geometry and response of the DGS cell which mentioned in Figure 2.18 incorporated with a microstrip line in CST.



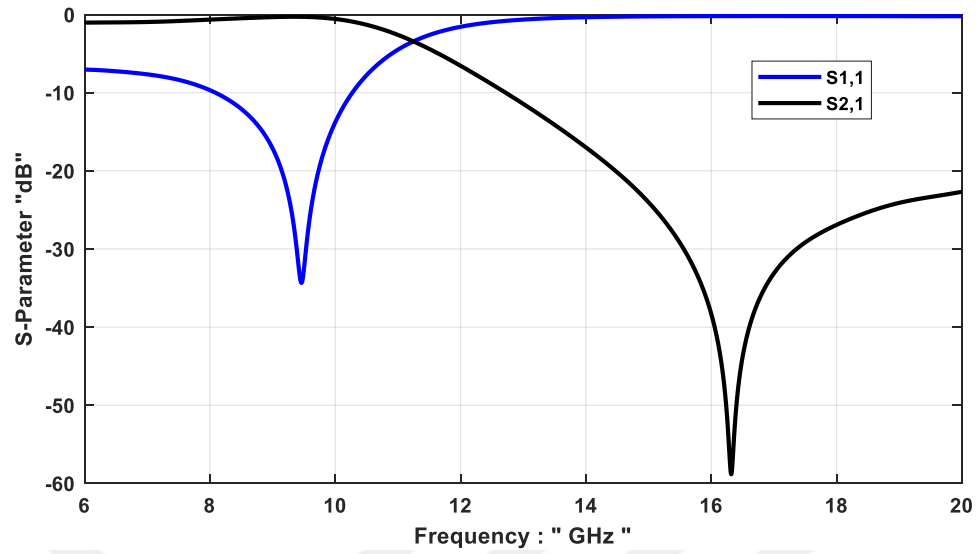
(a)



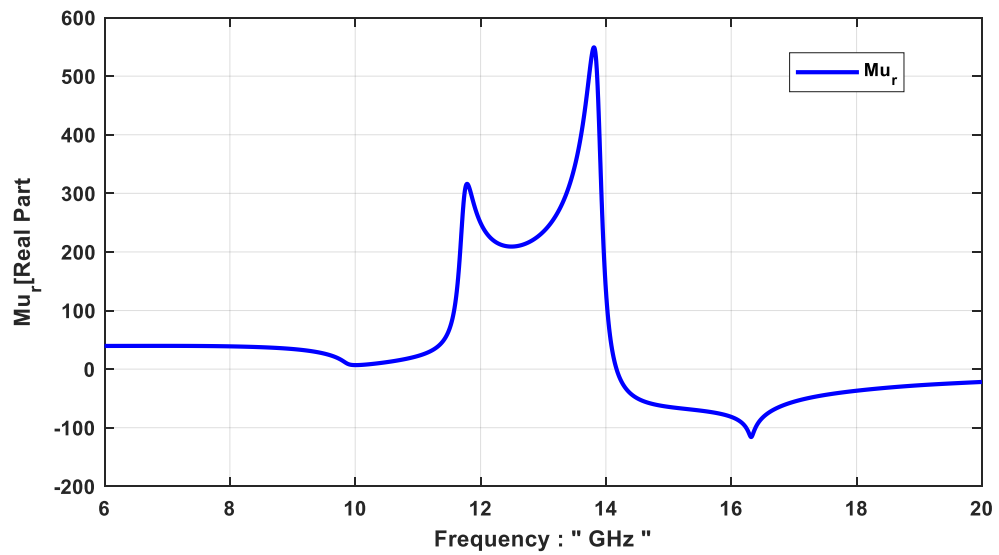
(b)



(c)



(d)



(f)

Figure 2. 19. Microstrip line loaded by DGS cell; **a)** 3-D structure; **b)** microstrip transmission line; **c)** DGS cell etched in the ground plane; **d)** S -parameters response; **f)** the real part of the permeability; “the yellow and orange indicates the copper and the white indicates the substrate’s insulation material”

2.3.1. DGS Applications in Microwave Circuit (L. H. Weng, 2008)

- Stopband or band gap Effects (demonstrated above)

DGS utilization in filter design is in two-level:

- One is directly using the band-gap effects of DGS in filter design.
- The other is using DGS in pre-constructed filters to improve performance, such as providing a sharper transition from passband to stopband also, suppressing higher modes or spurious responses, and hence result in improvements in the stopband characteristics.

- Slow-Wave Effect

Another property of DGSs is the slow-wave effect that arises due to the equivalent LC components. That comes from the fact that the transmission lines loaded with DGS possess higher impedance than the conventional ones, moreover, it has a higher slow-wave factor than the conventional ones. Slow-wave structures act to reduce the group velocity of a transmission line or increase its group delay compared to a normal "fast-wave" structure, so they are useful in shrinking the size of the microwave circuits.

- High Characteristic Impedance

The transmission lines loaded by DGS cells have much higher impedance than conventional lines. That is due to the presence of DGS cells in the ground of transmission lines leads to an increase in the equivalent inductance and reduces the equivalent capacitance of the lines, which results in an increment in the equivalent impedance of that lines. For example, the conventional microstrip line impedance realization is limited to, around $100 \sim 130 \Omega$, after loading a DGS cell on the ground plane it can be increased to more than 200Ω for the same line width which has a lower impedance. And this was demonstrated in a design of Unequal Wilkinson Power Divider conducted by (Oh et al., 2007), Figure 2.20 demonstrates layout of the 1:6 unequal Wilkinson power divider designed by adopting a microstrip line with a simple rectangular-shaped DGS cell to realize 208Ω of characteristic impedance. The 208Ω cannot be realized using a conventional microstrip transmission line.



Figure 2. 20. Picture of the fabricated 1:6 Wilkinson power divider with DGS designed by (Oh et al., 2007)

3. METHODOLOGY

In this chapter, we present in brief the overall design strategy of the filters and the procedure followed to suppress the spurious response and optimize the overall response of the desired response. Before starting the basic design stages, the flowchart presented below illustrates the overall design method followed during the design process. For this thesis' two compact SIW-DGS bandpass filters have been designed, the first one is having center frequency of 10 GHz with 4 GHz bandwidth, and fractional bandwidth of (FBW= 40%), while the second design is having center frequency of 6.175 GHz with 500 MHz bandwidth (FBW=8.3%) designed for Space Application. The printed circuit board (PCB) used in the design is Diclad 880 woven fiberglass/PTFE composite material having a dielectric constant of 2.2, with dissipation factor 0.0009 at 10 GHz, and a thickness of 0.508 mm. The overall design process was held using CST – full-wave simulator, time-domain solver, also verified by the frequency-domain solver. Followed by fabrication and measurement to analyze the performance of the designed filters. The flowchart shown in Figure 3.1. illustrate the overall design process of the thesis, while the flowchart in Figure 3.2. shows the filter design process, and in Figure 3.3, we can see in the flowchart the strategies followed to get the initial design of the DGS cells used in the filter design process. In the next chapter, the design process is presented in detail.

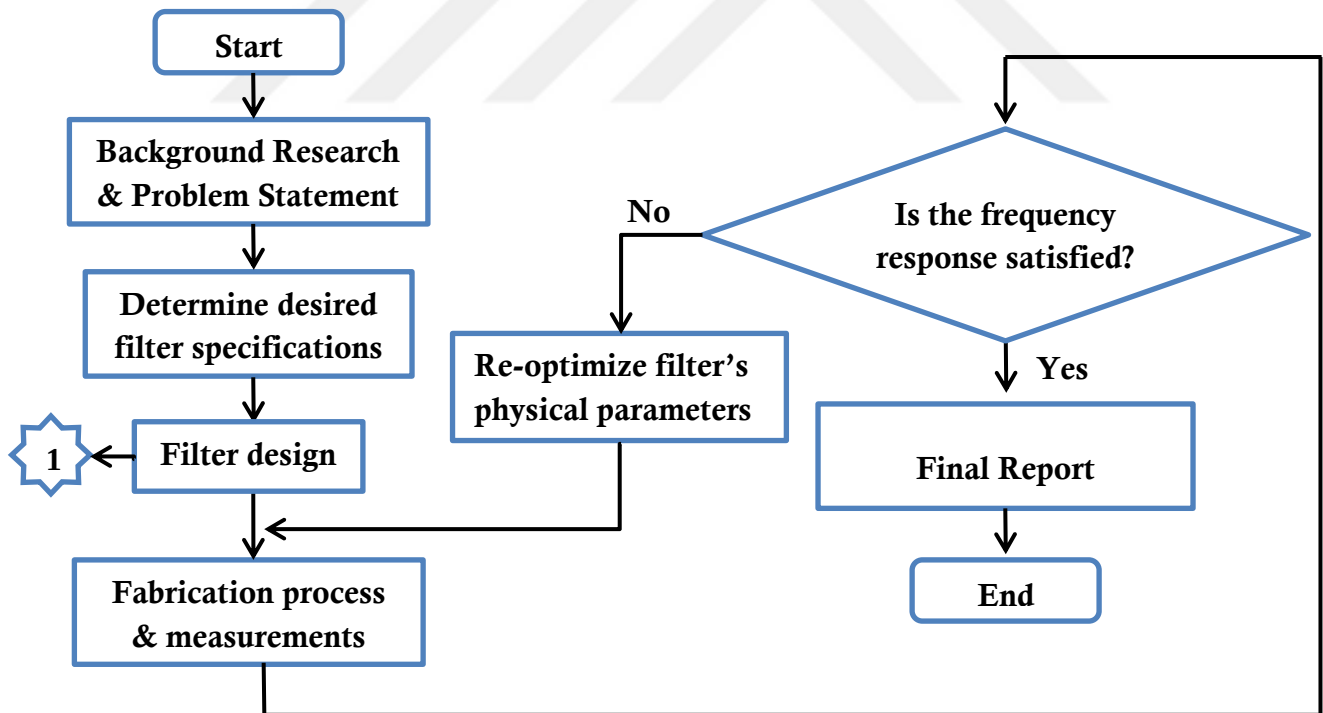


Figure 3. 1. Overall design process of the thesis

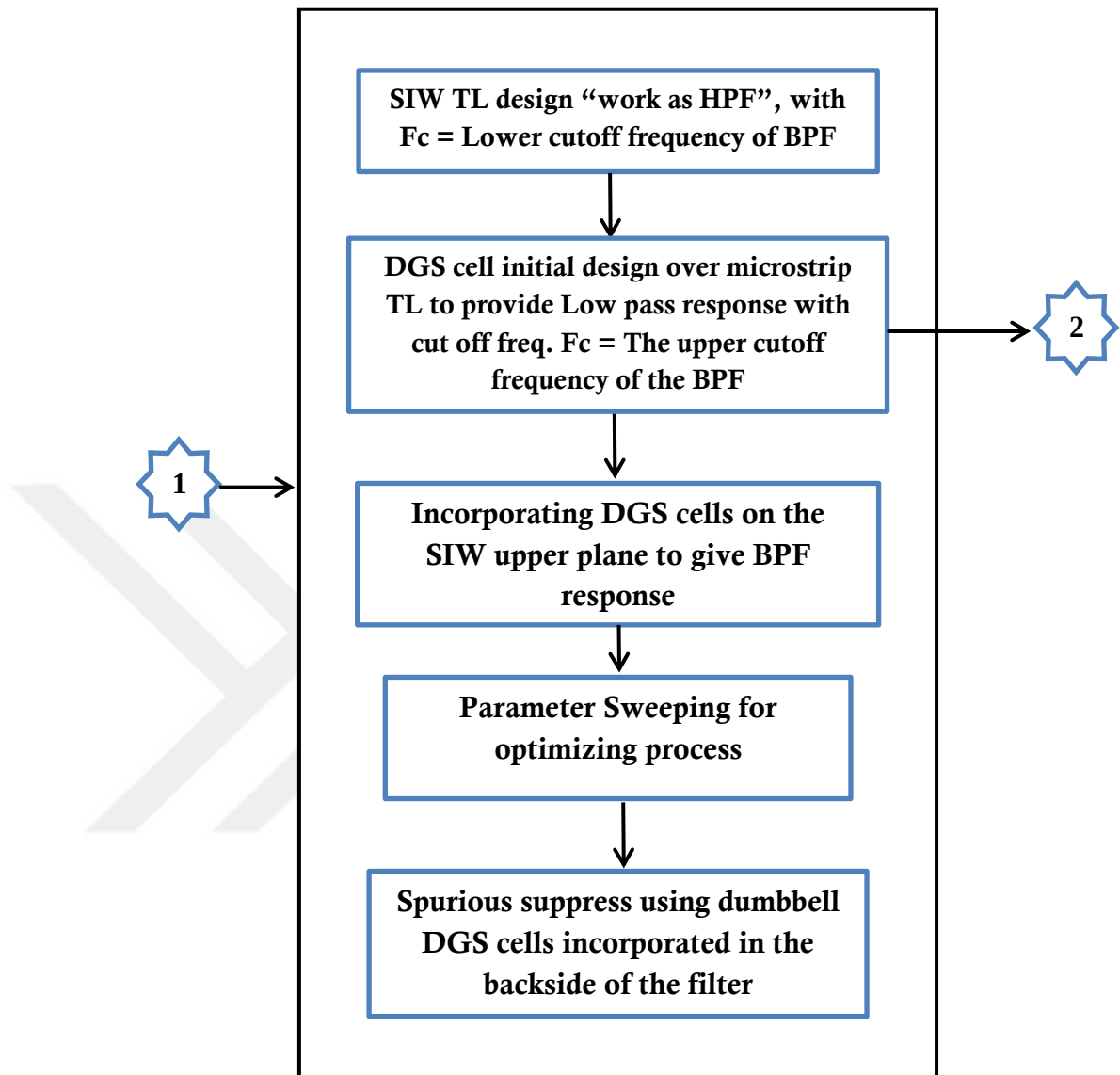


Figure 3. 2. Filter design process

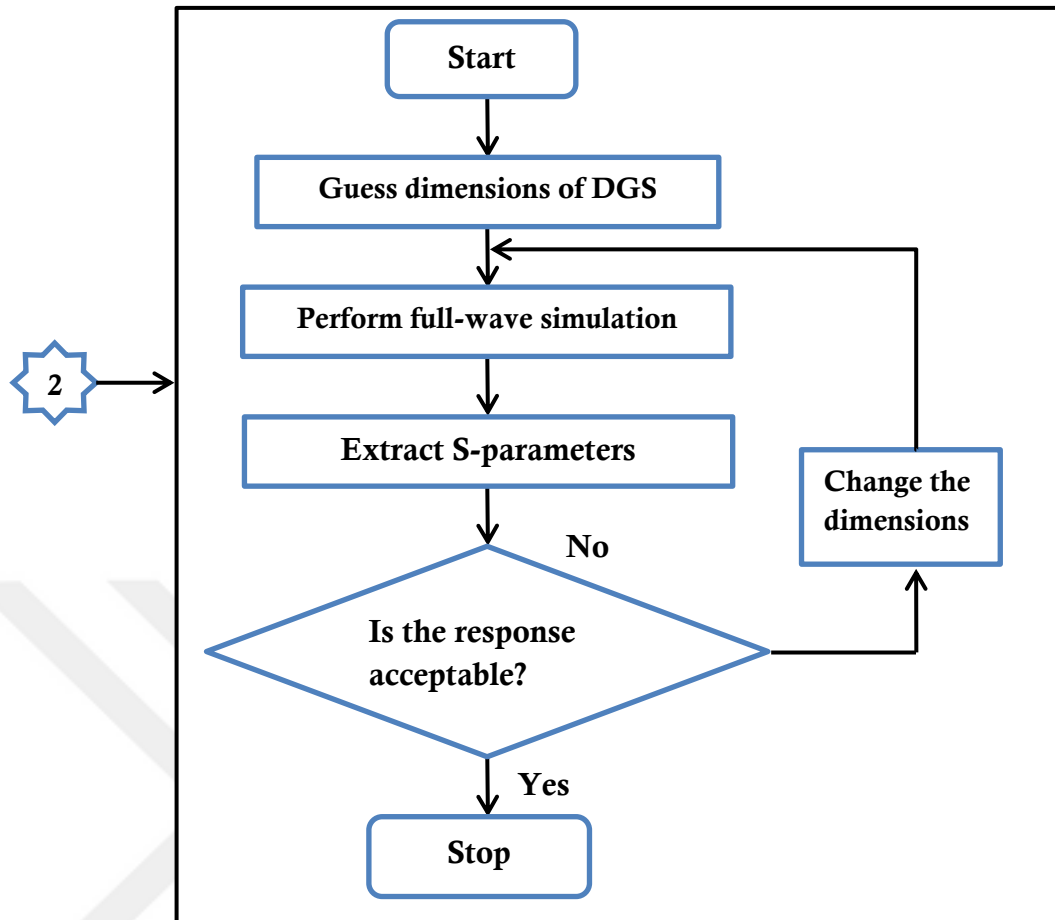


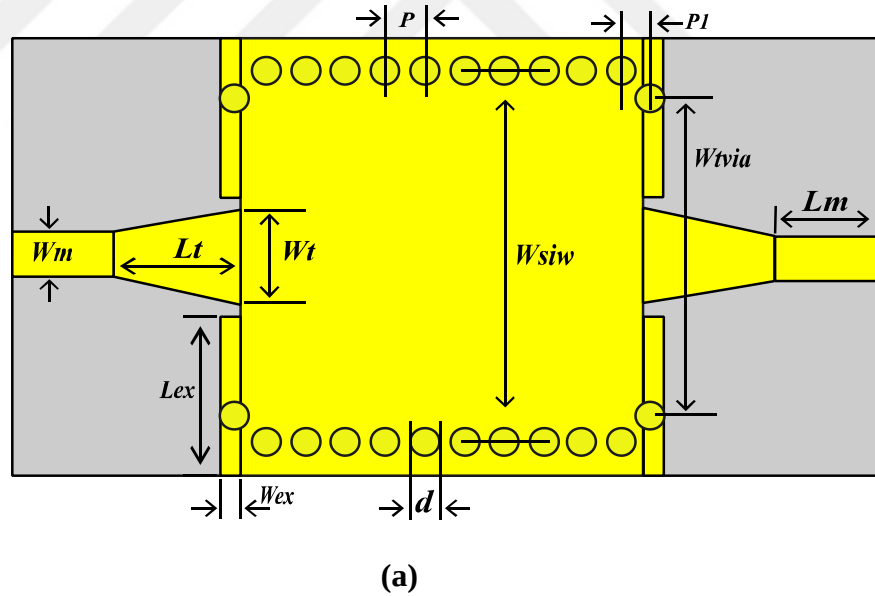
Figure 3. 3. Conventional design process of the DGS cells

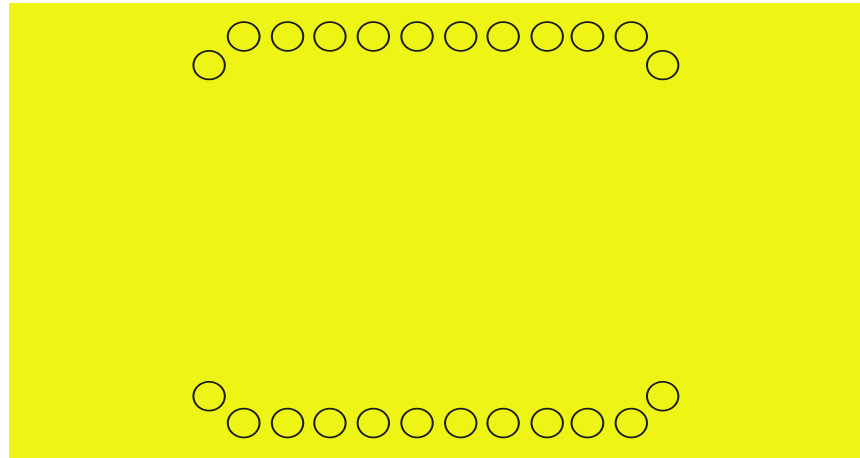
4. DESIGN PROCESS AND RESULTS

4.1. X– Band Filter

4.1.1. SIW & Taper transition design

As mentioned before the wideband X-band filter has a center frequency of 10 GHz with a fractional bandwidth of 40%, the design had built over Diclad 880 substrate that has a dielectric constant of $\epsilon_r = 2.2$, dissipation factor $\tan \delta = 0.0009$, and thickness of 0.508 mm. The first step in the design process is to calculate for SIW width that gives cutoff frequency corresponding to the lower cutoff frequency of the intended design equal to 8 GHz, the calculation was done using equations [2.23-2.28]. Then after SIW width calculation, the taper transition was designed, in which, the taper's initial design parameters were borrowed from (Kordiboroujeni & Bornemann, 2014), then optimization and modification were held, the taper geometry was modified to give less insertion loss and return loss, the combination of SIW with the taper transition gives the SIW the integrability with other microstrip circuits, this first part of the design process provides a high pass filter response with cutoff frequency at 8 GHz. The geometry of the design is illustrated in Figure 4.1.a, b, whereas the parameters corresponding to the dimensions of this configuration illustrated in Figure 4.1 are listed in Table 4.1. The simulation was run using CST MW studio, Figure 4.2 illustrates the resulting S-parameters.





(b)

Figure 4. 1. SIW transmission line; **a)** upper plane; **b)** bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”

Table 4. 1. Dimensions of X-band SIW & Taper which illustrated in Figure 4.1

Symbol	Value (mm)	Symbol	Value (mm)
Wm	1.919	Wsiw	13.9
Lm	3.6	d	1
Wt	4.187	P	1.9
Lt	3.6	Wtvia	12.4
Wex	0.5	P1	1
Lex	6.6	SubTh	0.508

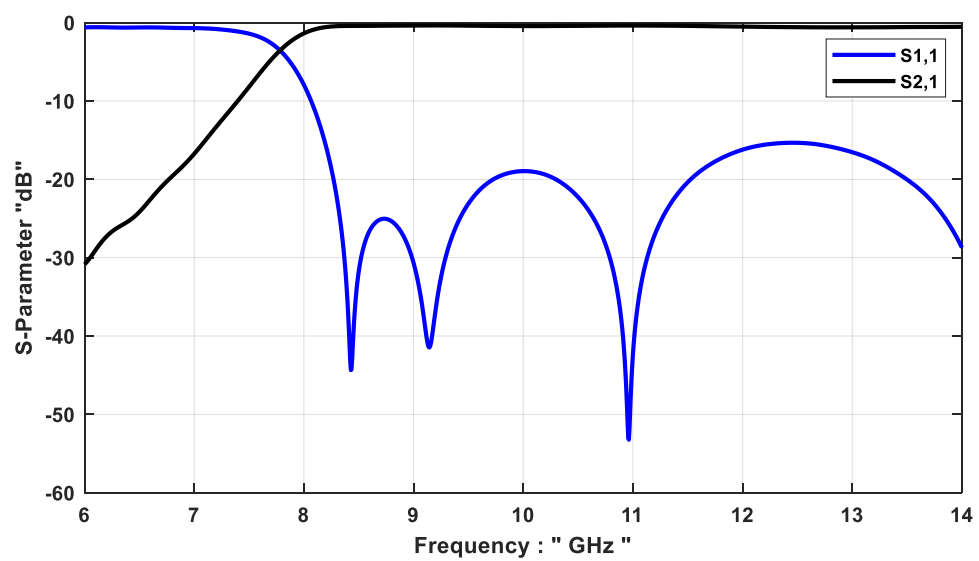


Figure 4. 2. The simulated S-parameters using CST

As we can notice from the S-parameters the insertion loss (S_{21}) around 0.4 dB, and return loss (S_{11}) better than -18 dB ranging from 8.3 GHz up to 11.5 GHz, in optimizing process we focused to obtain the minimum return loss (S_{11}) in the first half bandwidth of the intended design, since after loading the design with DGS cells the response corresponding to the upper half of the bandwidth will be affected the frequency response of the DGS cells.



4.1.2. DGS cell design

Moving to the next part of the design process, to yield the initial dimension of the DGS cell, the cell was designed using the conventional design process using the full-wave simulator CST MW studio. In order to save the time and conduct a fast parameter sweep until acquiring the initial dimensions of the cell, the DGS cell was modeled by a microstrip transmission line, where we etched the cell on the ground plane of a microstrip transmission line, as illustrated in Figure 4.3. Table 4.2 lists the dimensions of the cell shown in Figure 4.3, as for the simulation results, the s-parameters of related to this configuration are in Figure 4.4.

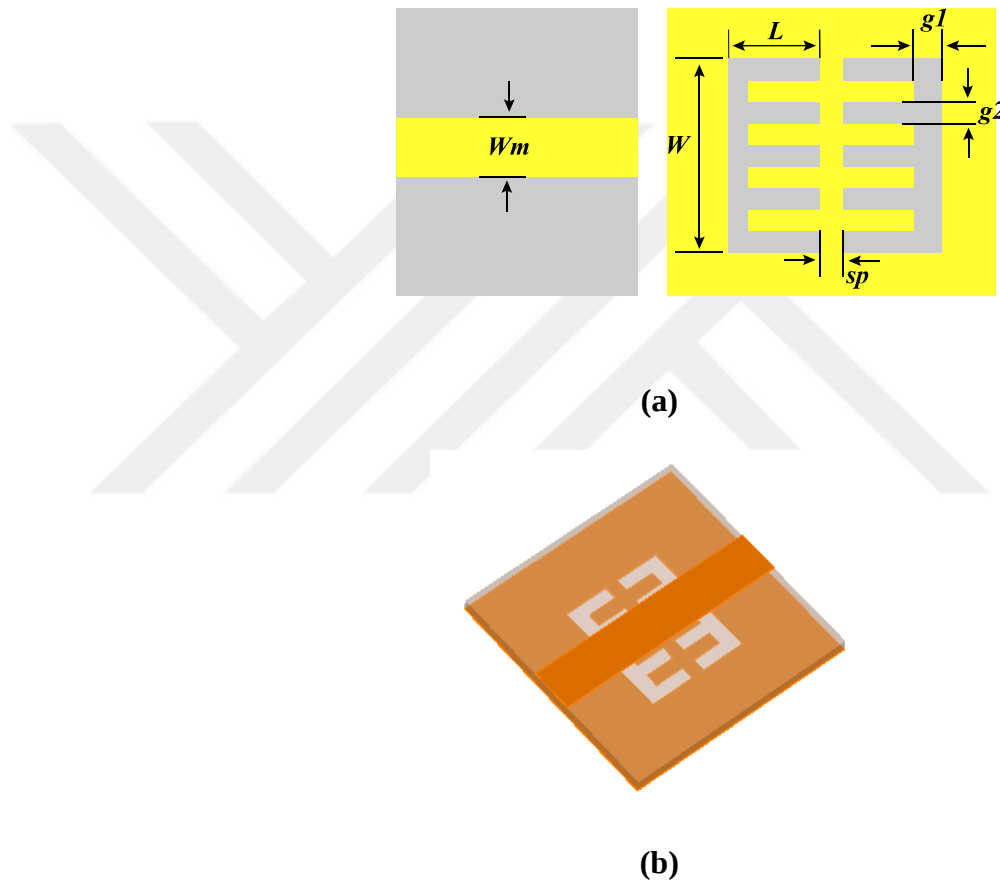
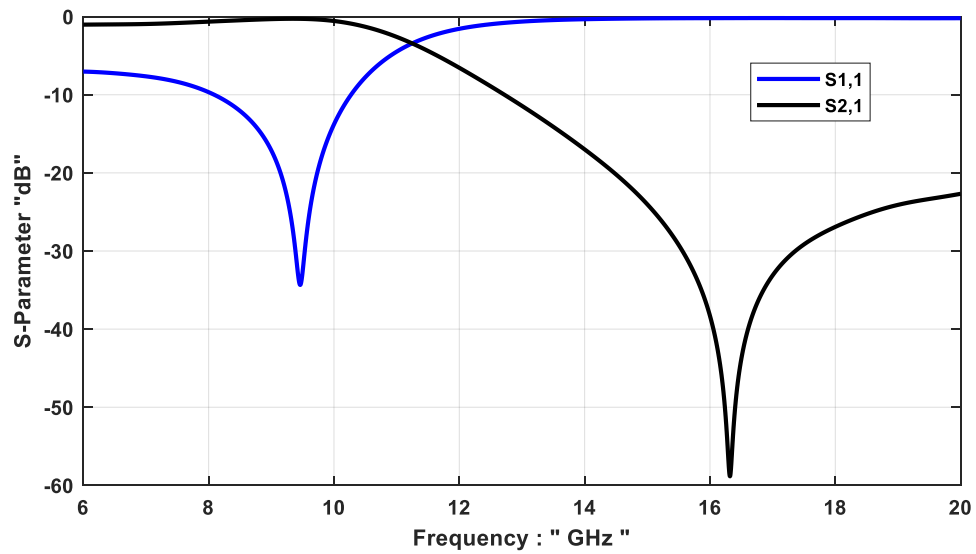


Figure 4. 3. DGS cell modeling using microstrip TL; **a)** Cell schematic; **b)** 3D-Structure; “the yellow and orange indicates the copper and the gray indicates the substrate’s insulation material”

Table 4. 2. DGS cell parameters

Symbol	Value (mm)	Symbol	Value (mm)
Wm	1.919	g1	0.51
L	1.75	g2	0.6
W	4.6	sp	0.5

**Figure 4. 4.** DGS cell simulated S-parameters using CST

We can see from Figure 4.4 that the DGS cell provides us band rejection in the region surrounding its resonance frequency. Etching the microstrip line with DGS cell produces a low pass filter as have here we have LPF with cut off frequency approximately 11 GHz and transmission zero located at 16.3 GHz. The next step in our work is to incorporate this cell with the SIW designed in section 4.1.1, to have a bandpass filter by combining high pass filter (SIW) with low pass filter (DGS) as we will see in the following sections.

4.1.3. 1st-Order SIW-DGS bandpass filter

The specifications of bandpass filter design which are determined before are as follow FBW 40%, Centre frequency of 10 GHz, with Diclad 880 laminate, to achieve this we start in this section with the first step by loading the SIW configuration by single DGS cell which has been modeled using a microstrip transmission line in section (4.1.2), here we are etching this DGS cell in the center of the upper plane of the SIW, which is designed before in section (4.1.1), this initial configuration represents the first step in the design process which is considered as 1st order BPF illustrated in Figure 4.5. the dimensions of the SIW and the DGS cell that building the configuration of the 1st order SIW-DGS BPF are the same as in sections (4.1.1, 4.1.2), the simulation was held using CST MW studio, Figure 4.6 presents the simulated S-parameters.

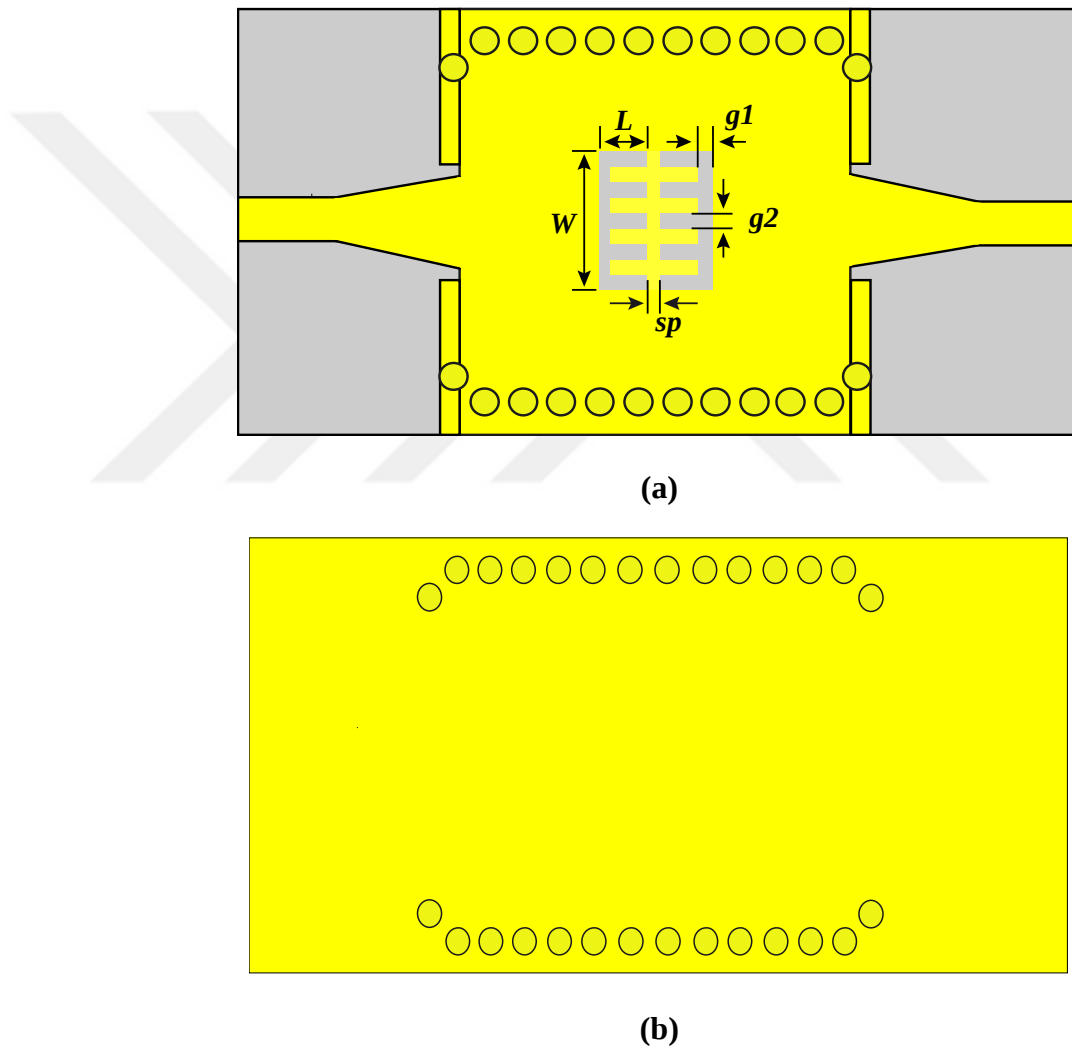


Figure 4. 5. 1st order SIW-DGS BPF; **a)** upper plane; **b)** bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”

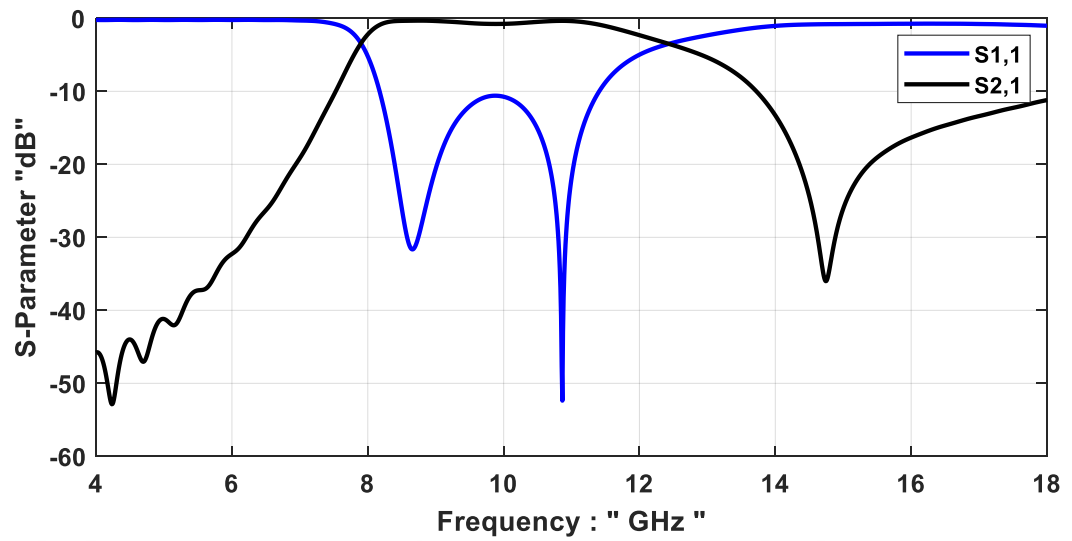


Figure 4. 6. S-parameters corresponding to the 1st-Order SIW-DGS BPF

We can notice from the simulated the S-parameters that combination SIW with DGS cell can produce BPF manner, we have insertion loss (S_{21}) 0.74 dB at the center frequency 10 GHz, and the return loss (S_{11}) in the passband better than -10 dB, -3db edges frequencies the lower at 7.9 GHz and the upper located at 12.26 GHz, for The rejection the transmission is better than -30 dB located at 14.7 GHz.

4.1.4. SIW-DGS 3rd order BPF

In this section, for the sake of better response, rejection, and sharper transition from passband to the stopband we increased the number of DGS cells in the SIW's upper plane to 3-cascaded cells which form a 3rd-order SIW-DGS bandpass filter. Three-cascaded identical DGS cells were added to provide a sharper transition from passband to stopband. We will notice from the simulation results of this configuration down below, the appearance of a spurious response in the upper stopband. The distance "s" that separate between the adjacent cells and the distance "sp" that separates between the two halves of the DGS cell, controls that spurious response, in order to minimize we have to assign these parameters carefully, to achieve the best performance in the passband, with a minimum spurious in the stopband. the geometry of the filter is shown in Figure 4.7, and the simulation results obtained using CST MW studio are presented in Figure 4.8. The values of "s" and "sp" parameters are 1 and 0.5 mm respectively.

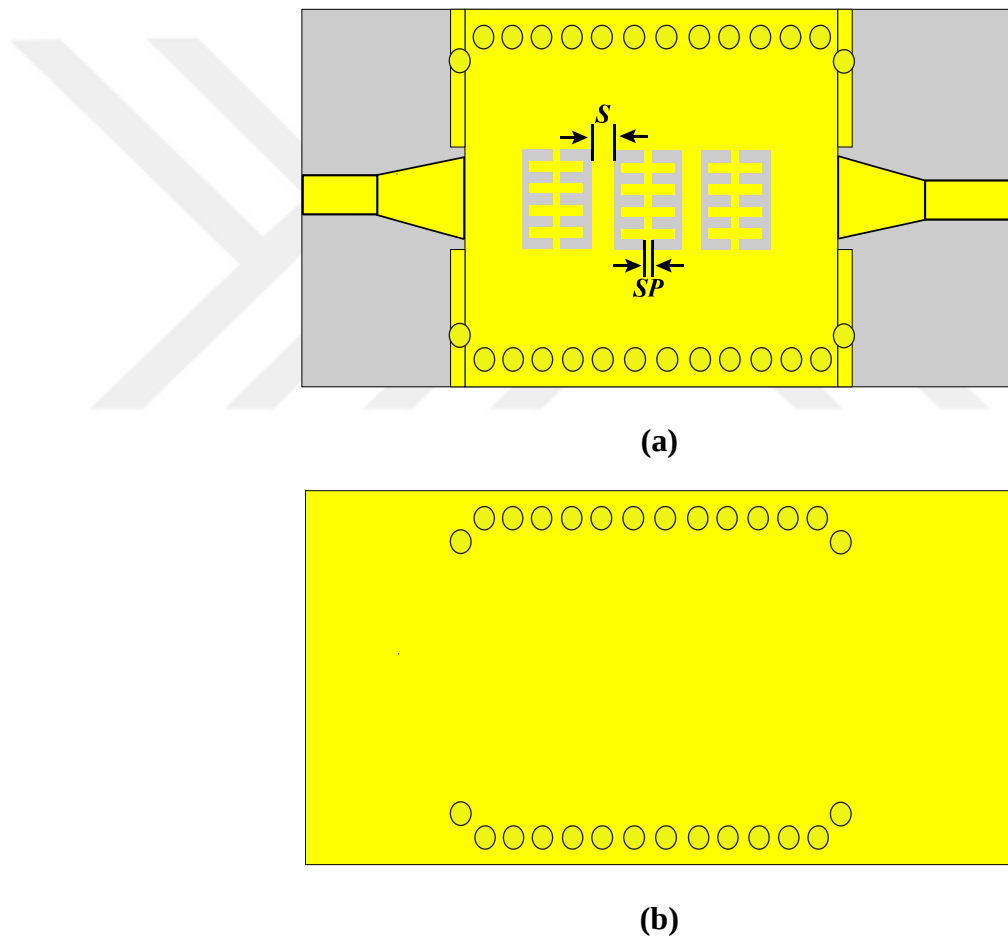


Figure 4. 7. 3rd order SIW-DGS BPF; **a)** upper plane; **b)** bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”

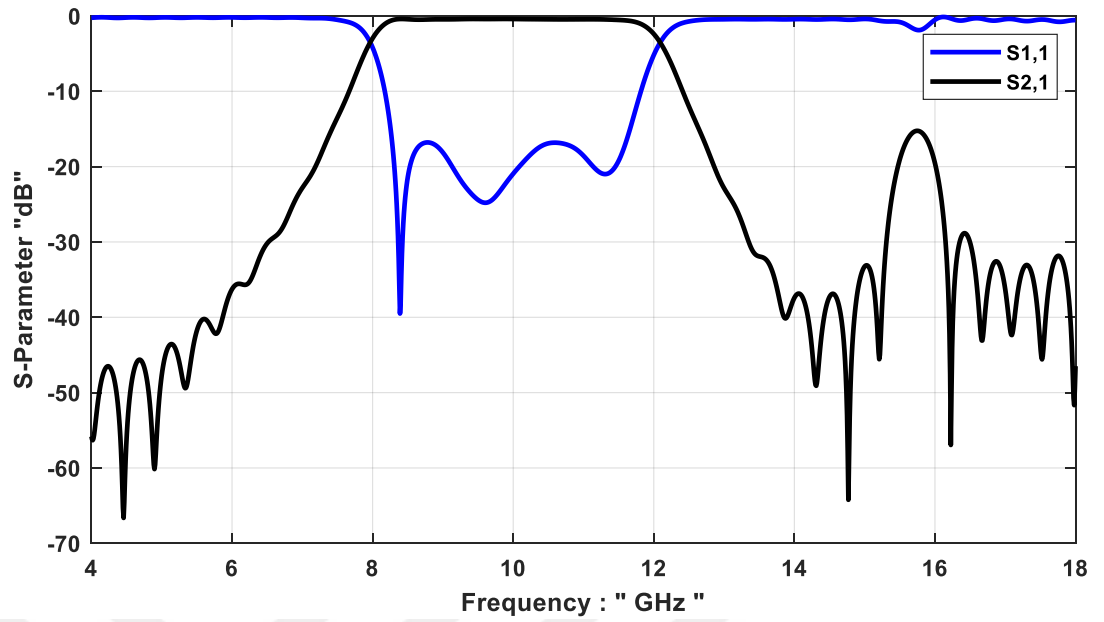


Figure 4. 8. S-parameters "simulation results" of the 3rd order SIW-DGS BPF

As we notice from the simulated S-parameters in Figure 4.8, we have insertion loss (S_{21}) 0.4 dB at the center frequency 10 GHz, and the return loss (S_{11}) better than -15.731 dB in the passband, the 3 dB upper edge of the filter at 12 GHz, and we have out of band rejection in the transmission coefficient (S_{21}) appear with a minimum rejection 15.2 dB ranging from 12.6 GHz until the spurious response appearance at around at 15.76 GHz.

4.1.5. SIW-DGS 3rd BPF – with one Dumbbell cell

In this section, in the sake for enhancing the overall response of the filter, we used one Dumbbell DGS cell etched in the center of the bottom plane located just beneath the center DGS cell located on the upper plane of the filter, as it represented in Figure 4.9, the dimensions of this dumbbell slot stated in Table 4.3, and simulation results using CST MW studio are shown in Figure 4.10.

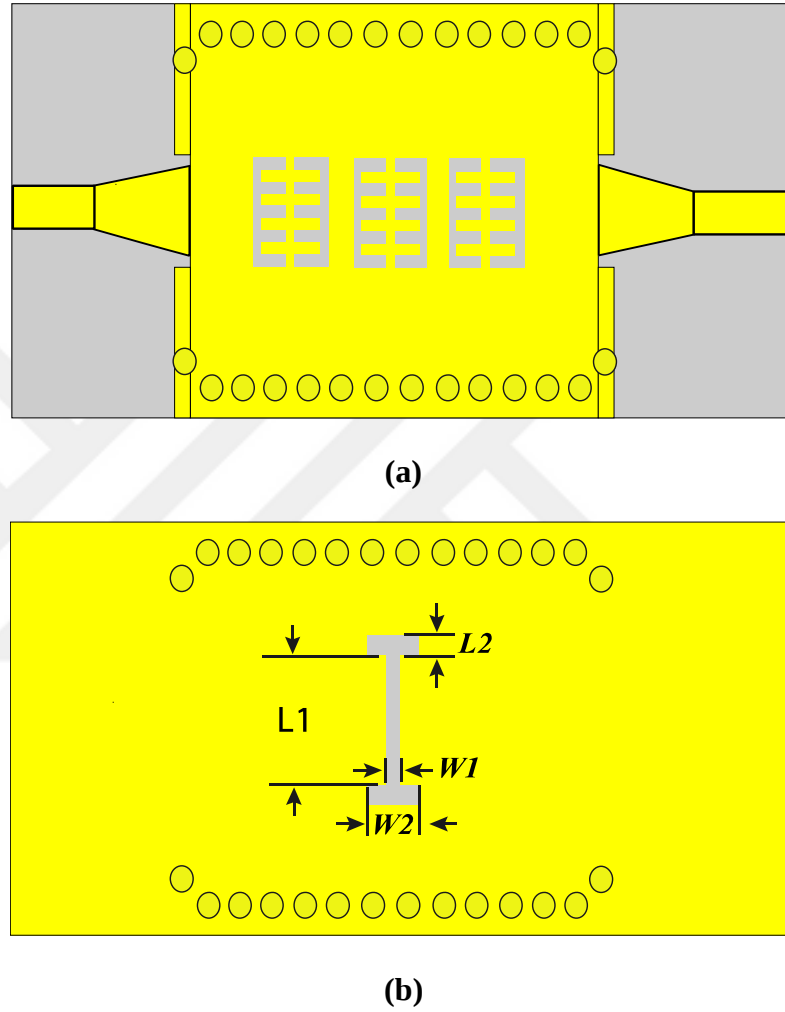
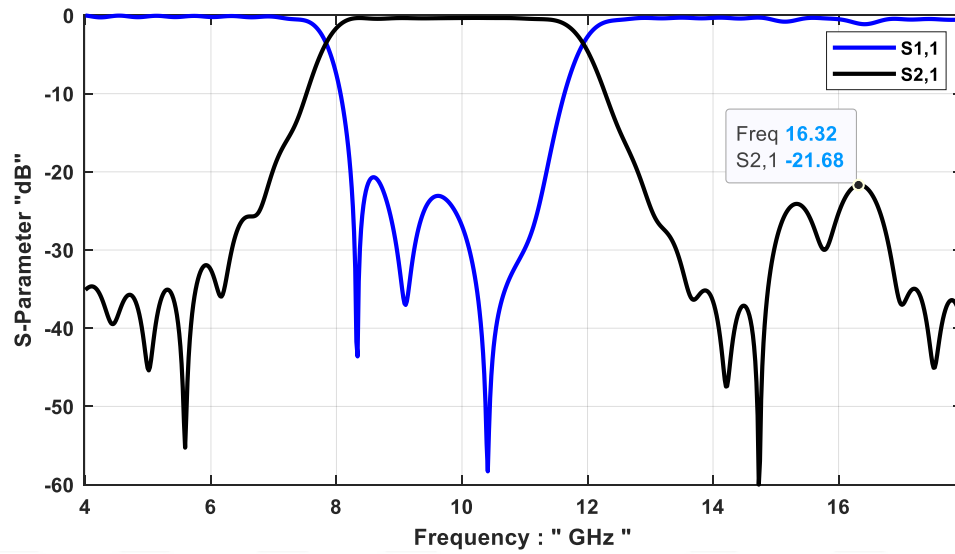


Figure 4. 9. 3rd order SIW-DGS BPF with one dumbbell in the bottom plane; **a)** upper plane; **b)** bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”

Table 4. 3. Center Dumbbell DGS dimensions

Symbol	Value (mm)	Symbol	Value (mm)
L1	4.7	L2	0.5
W1	0.35	W2	0.9

**Figure 4. 10.** 3rd order SIW-DGS BPF - with one Dumbbell cell simulated S-parameters

We can see from the simulated S-parameters that the loading procedure of the structure by dumbbell cell in the ground plane enhanced the coupling between the upper and bottom plane, which led to improvements in the response of the filter, in such that the insertion loss (S_{21}) now become around at center frequency of 10 GHz became to be 0.33 dB, while the return loss (S_{11}) became better than 20 dB in the passband range, also we got better rejection better than 21 dB ranging from 12.8 GHz up to the appearance of the spurious response at 16.34 GHz. Moreover, we can notice that the spurious response in this step has a maximum value of -21dB compared with -15.2 dB in the previous section.

4.1.6. SIW-DGS BPF - 3 Dumbbell DGSs etched in Ground plane

In this section, we sum up with the last step in the optimizing process of the X-band filter, in order to provide a further enhancement in the response of the filter and for better out-of-band rejection and more compact configuration, the first step done here was deleting four vias from each row to make the structure more compact, after that, we run the simulation and approximately there was no change in the response, after that, we have loaded the structure by extra two dumbbell cells, where they were etched in the ground plane, just beneath the microstrip ports lines. Figures 4.11.a, 4.11.b presents the upper and bottom plane of this configuration, while in Figures 4.12.a, 4.12.b we illustrate the simulated S-parameters and the Group-delay between inserted between the two ports of the filter. The dimensions of the port dumbbell tabulated in Table 4.4. The simulation process was held using CST MW studio time-domain solver.

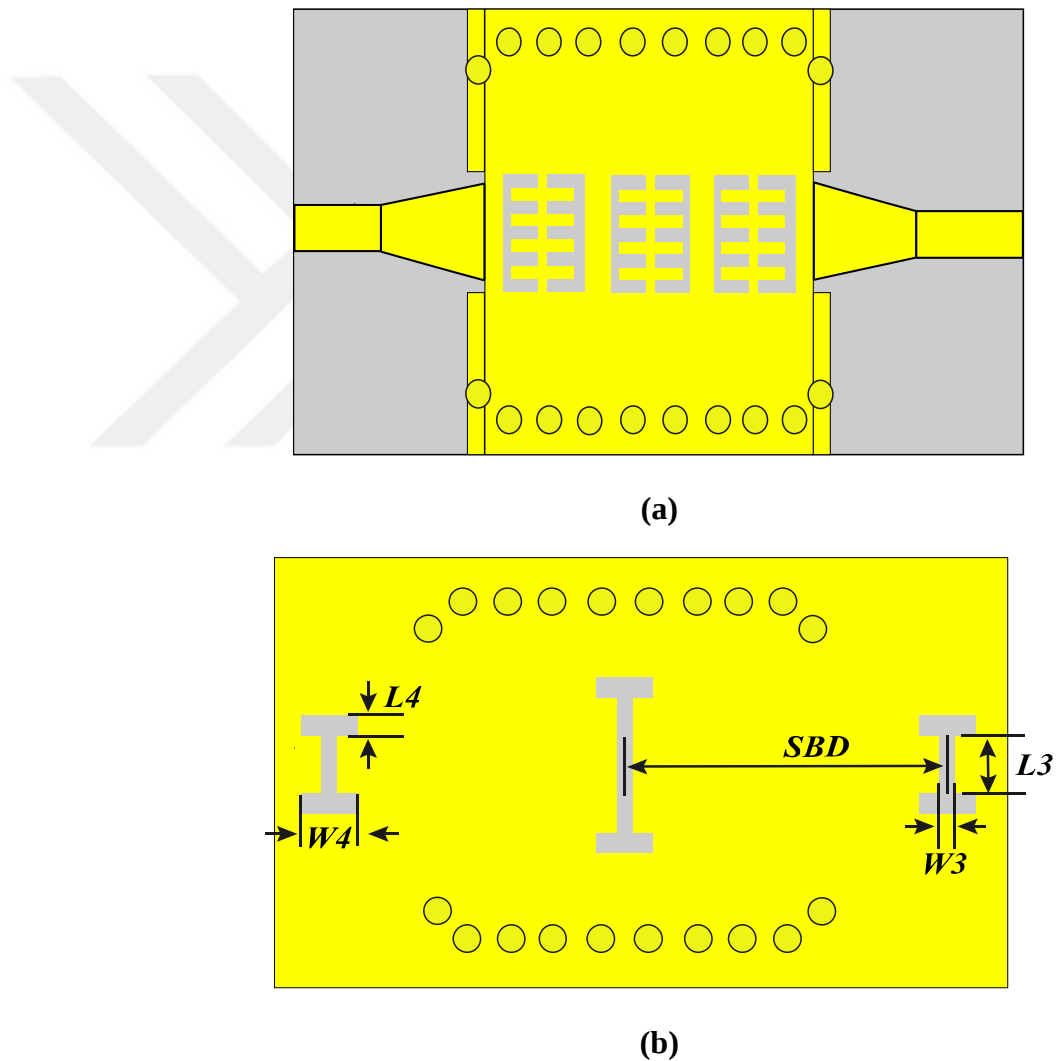
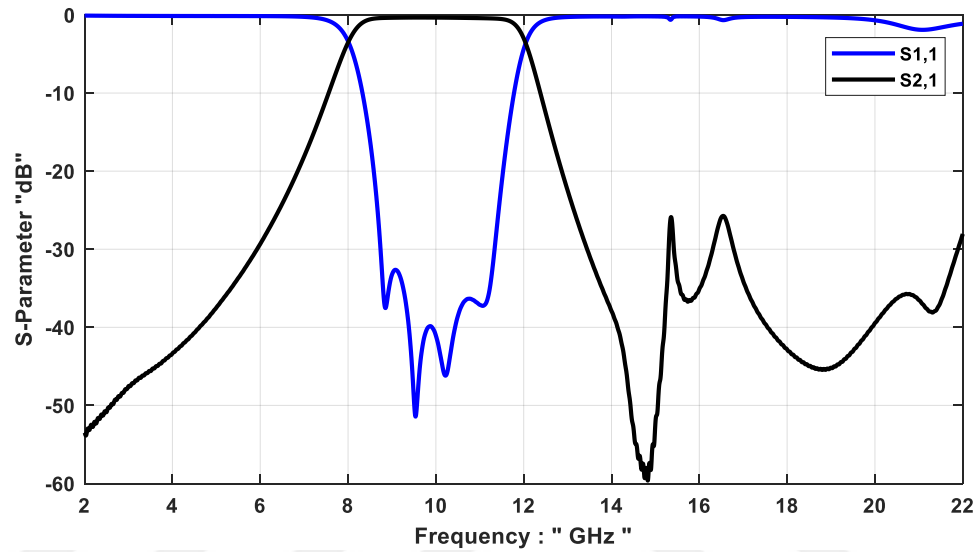
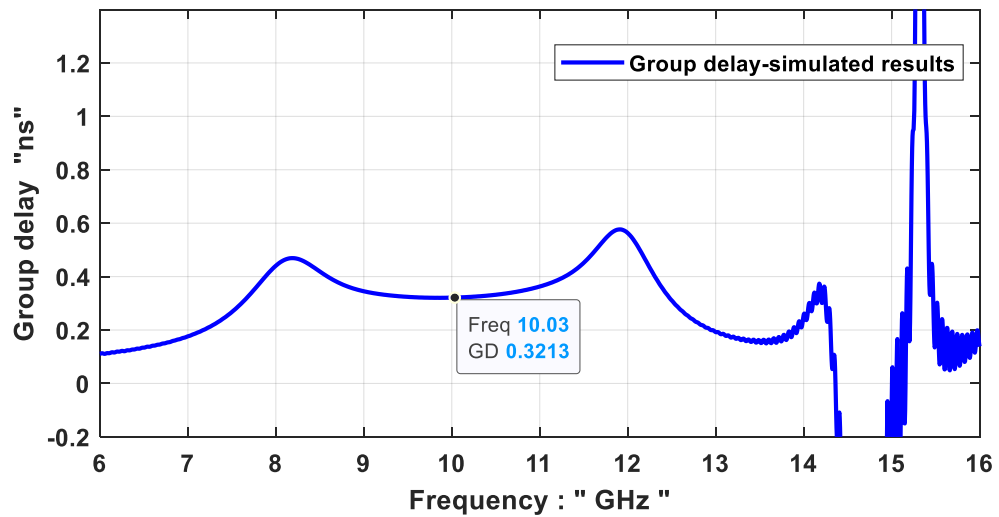


Figure 4. 11. 3rd order SIW-DGS BPF - with three dumbbells; **a)** Upper plane; **b)** Bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”

Table 4. 4. Port Dumbbells dimensions.

Symbol	Value (mm)	Symbol	Value (mm)
L3	1.7	W4	0.9
W3	0.3	SBD	12.5
L4	0.5		

**(a)****(b)****Figure 4. 12.** Simulation results of the 3rd order SIW-DGS BPF with three dumbbells; **a)** S-parameters; **b)** Group-delay

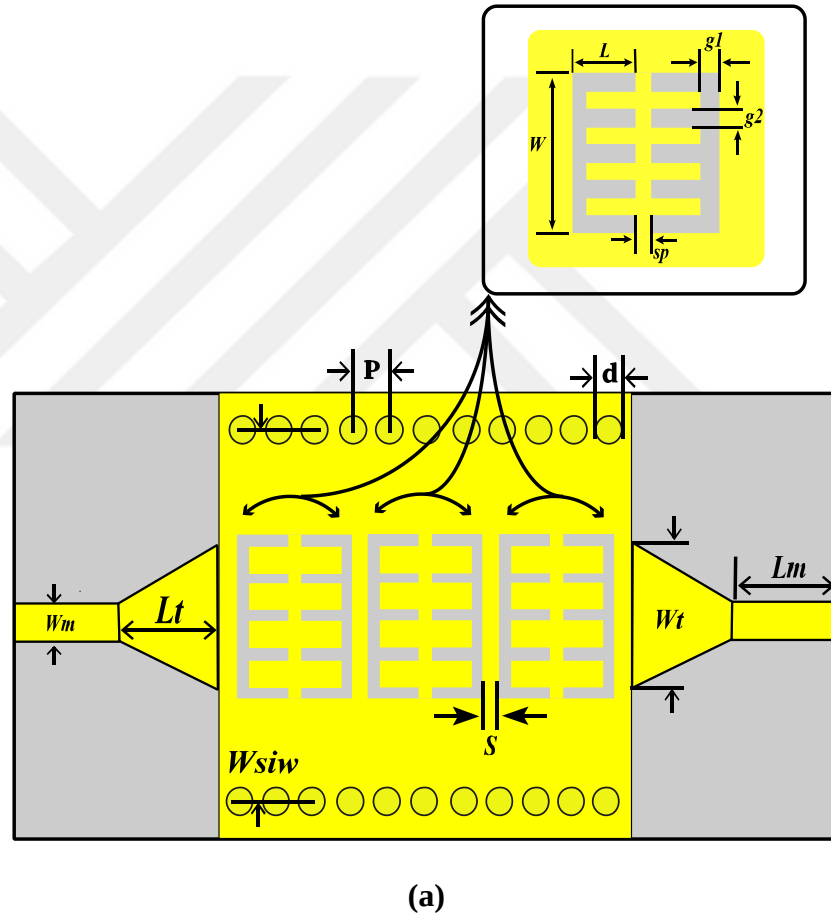
The dimensions stated in the Tables 4.1- 4.4 are the final values for the optimized result (by simulation), we can observe that after adding the dumbbells beneath the ports the response has significantly enhanced, from Figure 4.12. a, we can see approximately 0.35 dB insertion loss (S_{21}) at the center frequency of 10 GHz and the return loss (S_{11}) better than -32 dB in the passband, the transmission coefficient almost flat in the passband, regarding the upper rejection we have a wide out-of-band rejection of 25 dB ranging from 13 GHz to further than 22 GHz, the spurious response has been significantly reduced to lower than 25 dB at (15.3 and 16.5) GHz, and a transmission zero of -58dB rejection at 14.7 GHz, as for the group delay of between filter ports it is illustrated in Figure 4.12.b we can notice that it is approximately flat within the passband, with a delay about 0.32 ns at the center frequency of 10 GHz, and the variation is quite smooth for a microwave filter. Finally, regarding the size of the filter, the filter preserves the miniaturization property with a total size of $17.7 \times 30 \text{ mm}^2$.

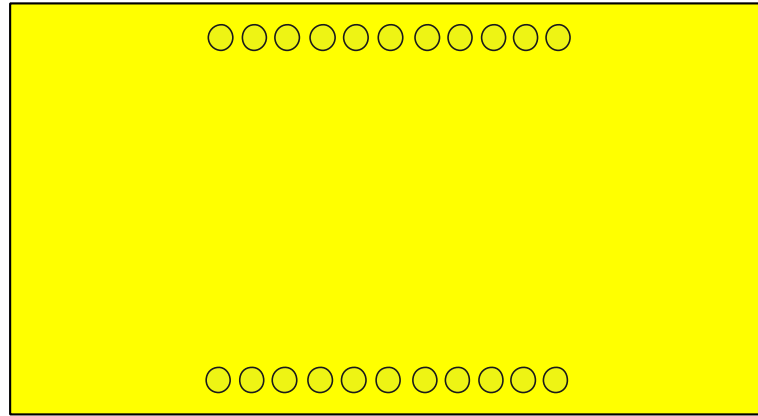


4.2. C– Band Filter

Our goal is to present a simple compact planer design of a bandpass filter having specifications as follows, a center frequency of 6.175 GHz with 500 MHz bandwidth (FBW=8.3%), to emulate the response of non-planer filter introduced by (Tomassoni et al.,2017).

Our proposed filter forms 3rd order SIW-DGS C-Band filter, combining SIW structure loaded with 3 cascaded DGS cell, the same cell structure used in section 4.1.2, with the same substrate. the resulting filter structure is presented in Figures 4.13 a, b, and its dimensions are tabulated in Table 4.5. The filter modeled using CST MW full-wave simulator, and simulation conducted using the time-domain solver. The simulation result: S-parameters, and group delay illustrated in Figure 4.14. a, and Figure 4.14. b.



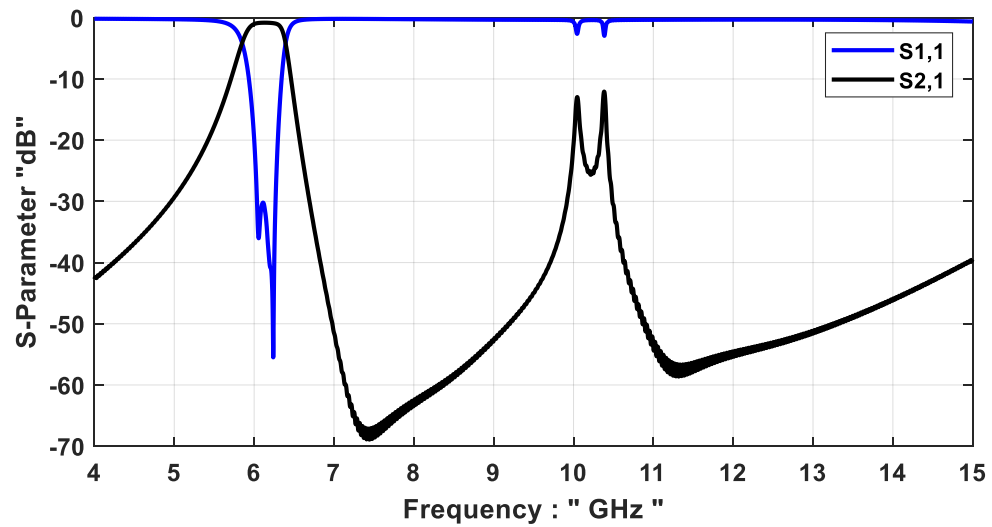


(b)

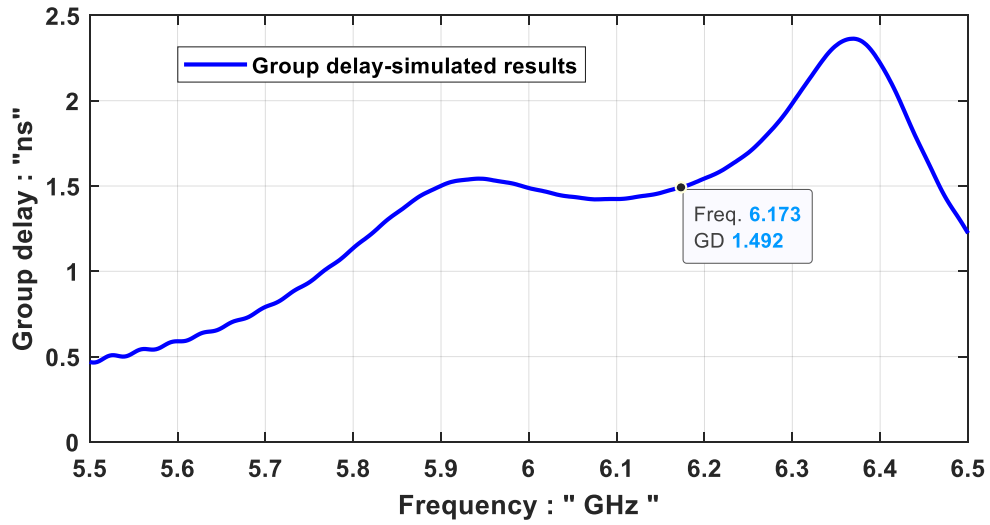
Figure 4. 13. 3rd order SIW-DGS C-Band BPF; **a)** Upper plane; **b)** Bottom plane; “the yellow indicates the copper and the gray indicates the substrate’s insulation material”

Table 4. 5. C-Band filter dimensions

Symbol	Value (mm)	Symbol	Value (mm)
Wm	1.519	SP	0.8
Lm	3.8	S	0.8
Wt	7.587	L	2.5
Lt	4.7	W	8.47
Wsiw	16.25	g1	0.5
d	1	g2	0.6
P	1.9	SubTh	0.508



(a)

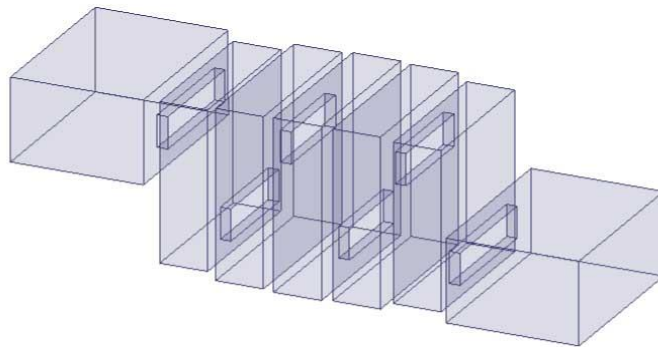


(b)

Figure 4. 14. 3rd order SIW-DGS C-Band BPF; **a)** S-parameters; **b)** Group-delay

The interface transition between SIW and microstrip used in the design is a taper transition that has been designed using the full-wave simulator, the filter is a simple combination of SIW loaded with 3 cascaded DGS cells. As we can observe from S-parameters that filter has 3-dB bandwidth of 500 MHz ranging from (5.925-6.425) GHz, the insertion loss (S_{21}) at the center frequency of 6.175 GHz around 0.80 dB, and return loss (S_{11}) in the passband better than 30 dB, regarding the out-of-band rejection we can notice the sharp transition from the upper 3-dB edge frequency at 6.425 GHz to the zero transmission of -68 dB at 7.4 GHz, the spurious response is appearing at 10.05 and 10.38 GHz lower than -12.2 dB. As for the filter's size, the filter preserves the compactness property with a total size of 20*37.9 mm².

The 3rd order SIW-DGS C-Band filter offers good performance when compared with the non-planar 5th order filter C-band realized by using single-mode TM resonators introduced by (Tomassoni et al.,2017), illustrated down below in Figure 4.15.



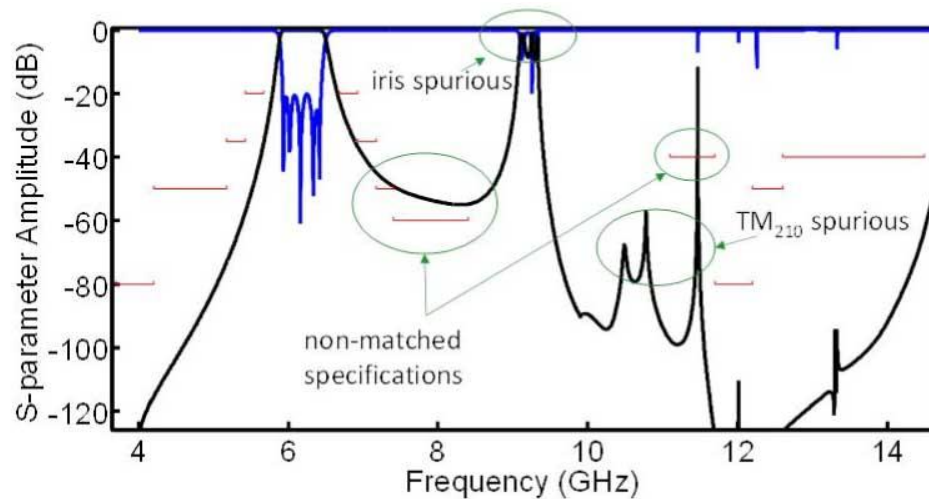


Figure 4. 15. 5th order filter in C-band realized by using single-mode TM resonators (top) and relevant full-wave simulated response simulate with HFSS (bottom) (Tomassoni et al., 2017)

Figure 4.15 above illustrate the 5th order C-band filter realized using single-mode TM resonators, and the relevant simulated response, if we want to make a comparison between the high-performance non-planer filter and our planer one realized using SIW-DGS, we can say the latter present good performance. Both suffer from a spurious response. The non-planer design insertion loss in first is about 0.2 dB while our design has insertion loss around 0.81 dB in the passband, as we see from the simulated result for both filters, we can conclude that the SIW performance is better against spurious response.

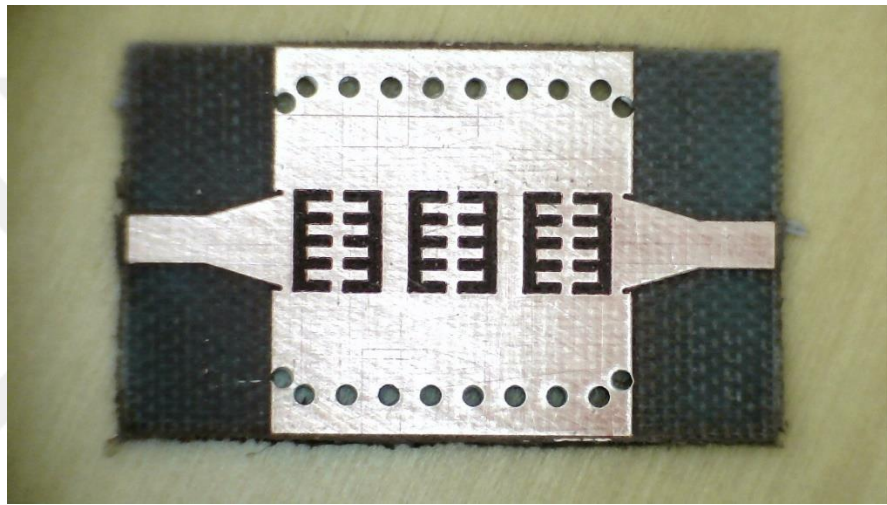
In (Tomassoni et al., 2017) they overcome the spurious response, by incorporating two bandpass filter to form 9th order one (filter + pre-filter) to have upper stopband free of spurious response, another way to suppress the spurious, is to accompany the output terminal of the bandpass filter with a Low pass filter, but this way make the design bulky.

For our design there is two way for enhancing the response and provide better suppression to the spurious the first, we can change the transition as we did with the X-band filter this lead to reduce the insertion loss and provide better return loss. The second, for suppressing process of the spurious responses, is to support the design with extra DGS cells in the ground plane that can provide better suppression over a wider range, Also, we can re-tune the dimensions of the DGS cells to reduce the spurious responses and without affecting the passband performance, especially tuning the parameters affecting the spurious significantly, which are the distance separating between the two halves of the cell “SP”, and the distance separating between the adjacent cells “S”.

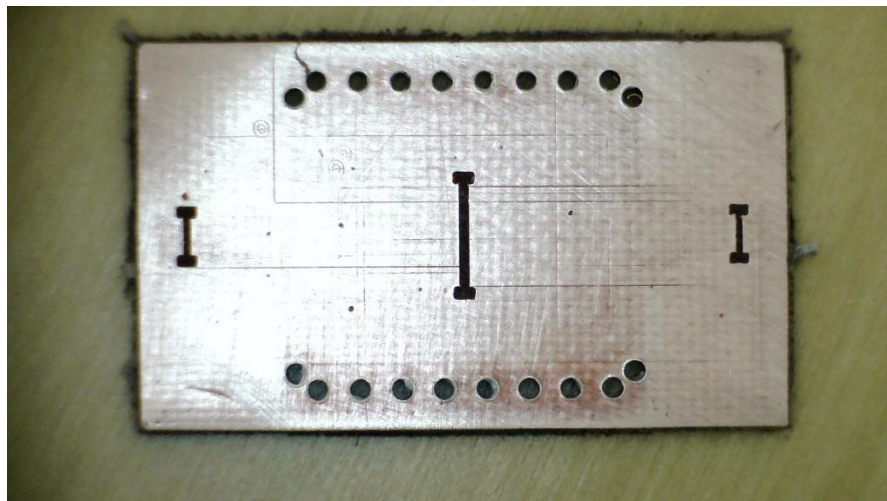
5. MEASUREMENTS

In this chapter, we discuss the fabricated sample and related measurements. To conduct the measurements of the fabricated filters, Ceyear 3672c-S Vector Network Analyzer is used. Calibrated by full two-port SOLT calibration. In the subsections below we illustrate the measurement process and the problems we faced during measurements.

The fabricated sample of X-band filter is shown in Figure 5.1, during the fabrication the milling machine faced difficulties with Rogers DiClad 880 laminates based on woven fiberglass reinforced, PTFE-based composites substrate, which cause it to be not rigid enough, has flexibility, and a degree of softness, so the LPKF S104 milling machine cannot etch the copper at high degree of efficiency at edges during processing, as we can notice through the fabricated sample in Figure 5.1 below.



(a)



(b)

Figure 5. 1. Fabricated X-Band filter; **a)** Upper plane; **b)** Bottom plane

The SMA connectors that were available with us during the measurements are shown in Figure 5.2 also include their related dimension. The distance between the inner conductor pin and the ground legs about 1.4 mm which does not match the laminate thickness of 0.508 mm, which led to confusion in the installation and so the measured results.

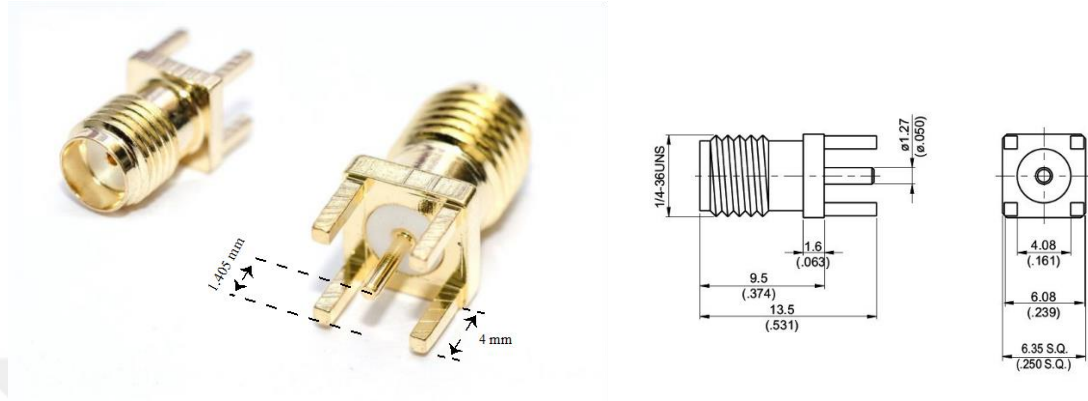
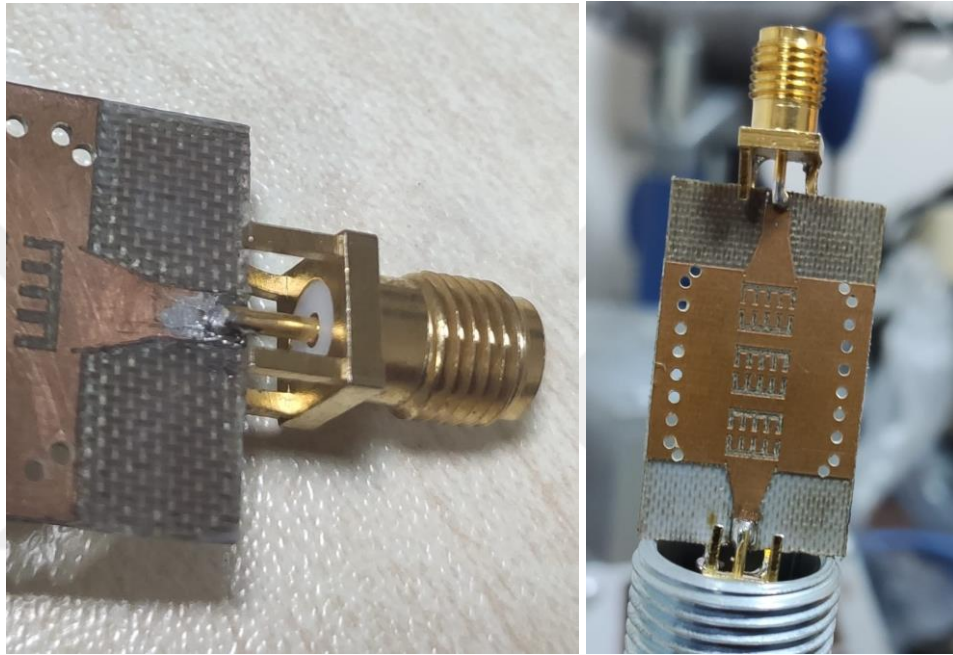


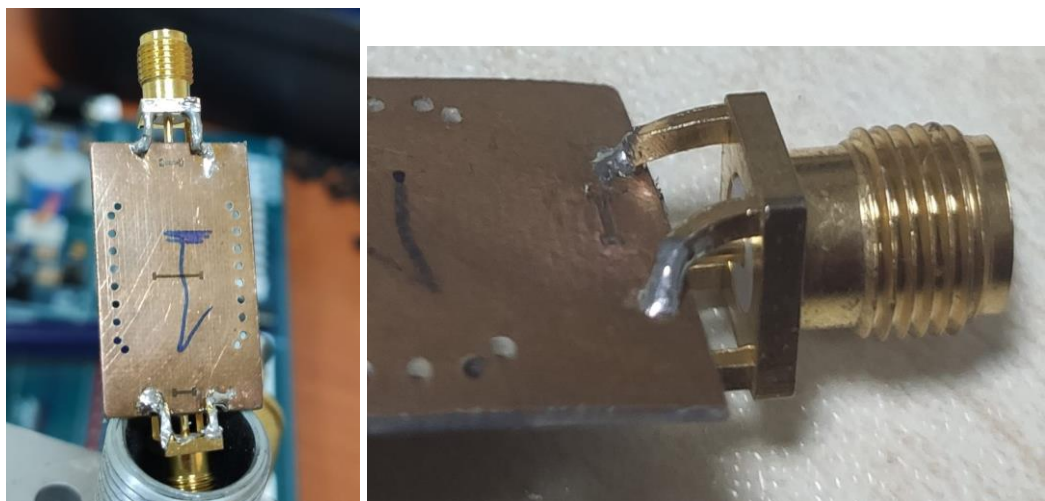
Figure 5. 2. SMA connector and its corresponding dimensions

5.1. X-Band measurements

The first installation of the SMA connector illustrated in Figure 5.3, as we can notice we tried to avoid any problem that could result due to coupling between the SMA legs and the dumbbell DGS cells etched beneath the filter ports, so at the ground plane we tried to bend the connector legs to push them away as much as possible from the dumbbells to avoid any coupling that could disturb the measurements. Also, the inner pin length of the SMA connector is greater than the filter's microstrip ports length, so we soldered the connector in such a way the SMA pin stick with the first half of the microstrip ports as we see in Figure 5.3.



(a)



(b)

Figure 5. 3. Bad installation of SMA connector; **a)** upper face; **b)** bottom face

After this installation of the connector as mentioned above, we put the filter under test to conduct the measurements, we got very bad results as shown in Figure 5.4.

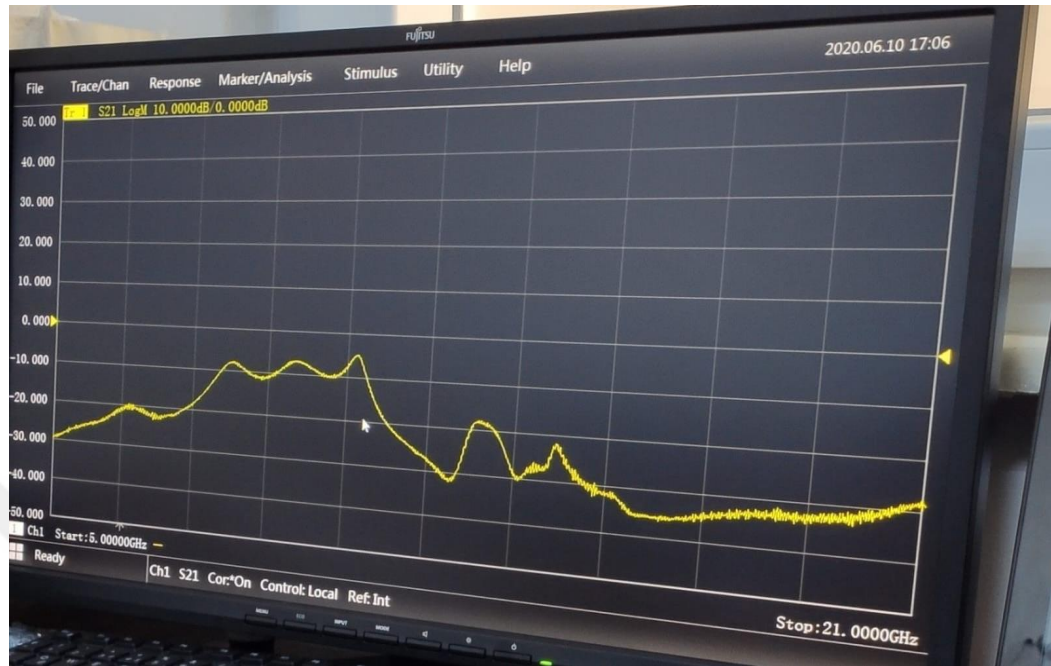


Figure 5. 4. Measured S_{21} response

A comparison between the ideal simulation results and the measured results are shown in Figure 5.5 below, we can see that the measured results are not acceptable. So, to investigate the reasons that led to disturbing the measurements, we modeled the SMA connector alongside the filter in the simulator to see its effect, at the same form we installed during the measurements, as shown in Figure 5.6, a comparison between the simulation results taking into account SMA effects and the measured results illustrated in Figure 5.7.

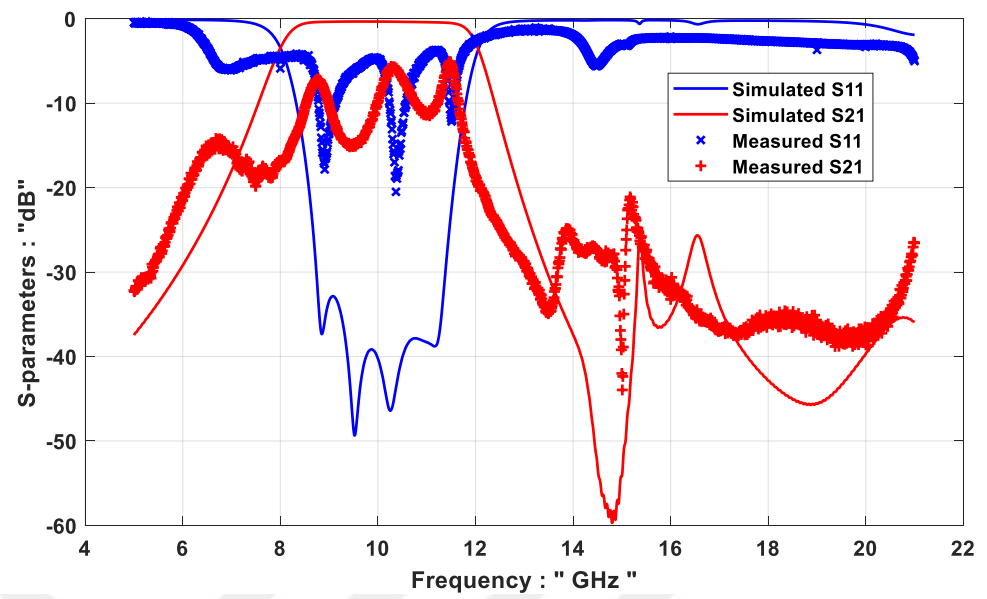
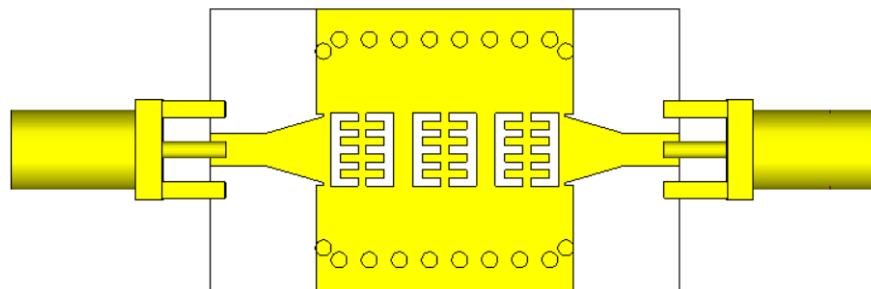
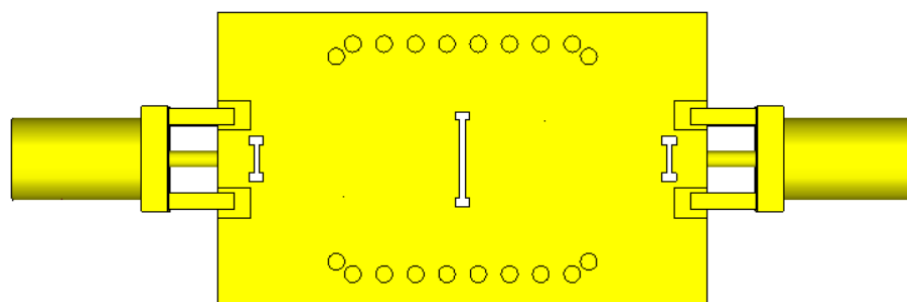


Figure 5. 5. Ideal simulation results vs. measurement results



(a)



(b)



(c)

Figure 5. 6. SMA inclusion in the simulation; **a)** upper view; **b)** bottom view; **c)** side view

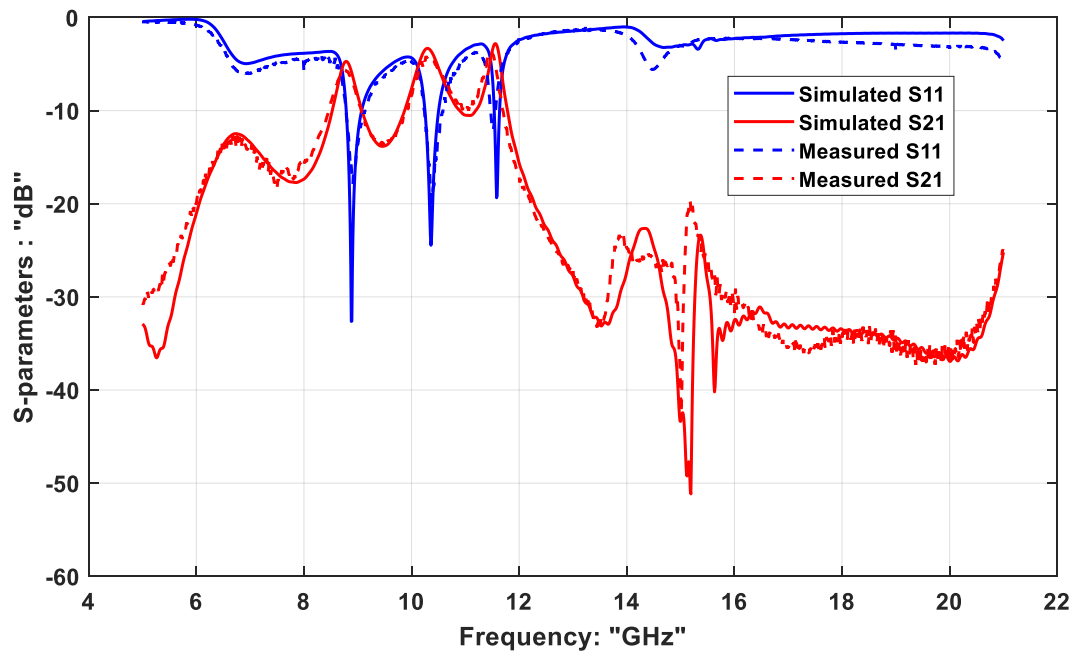
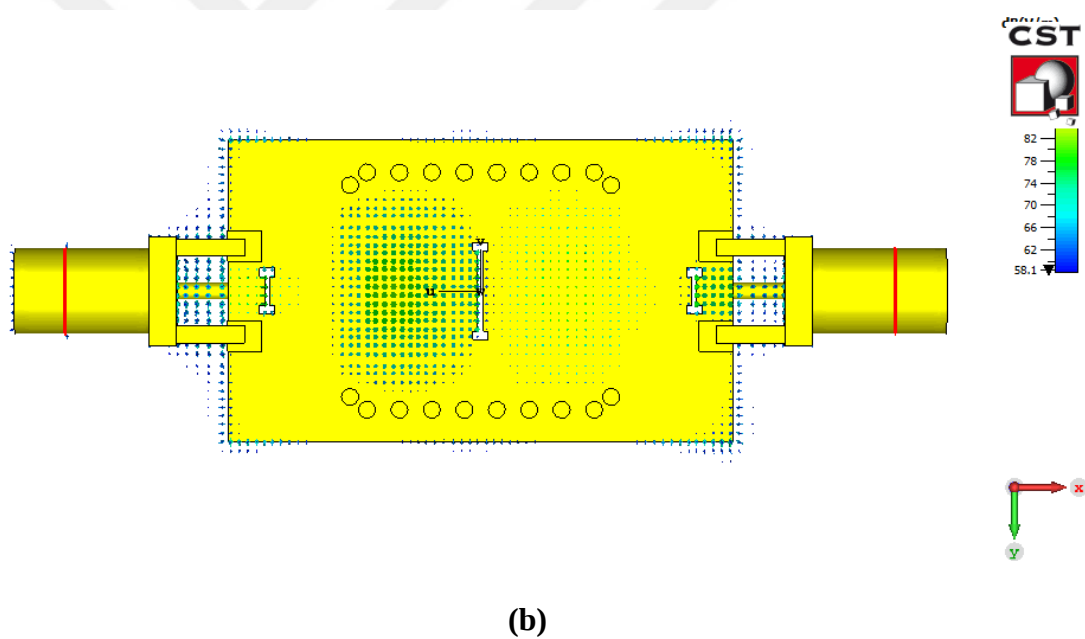
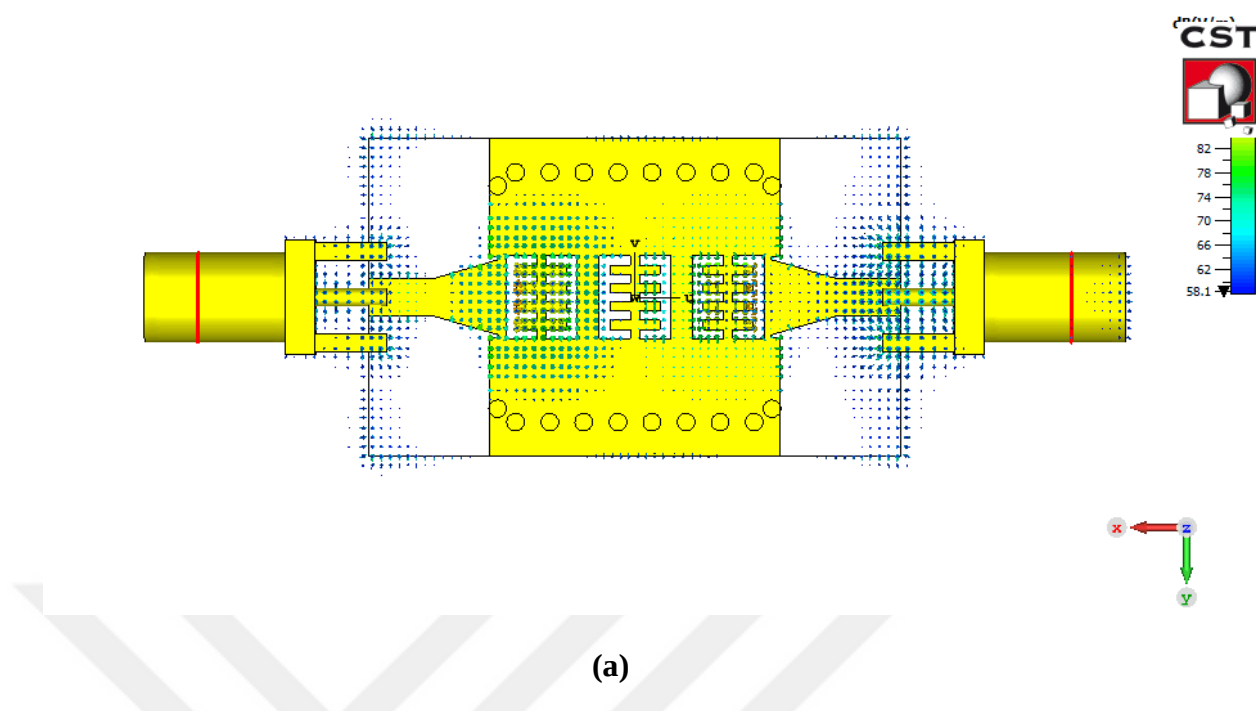
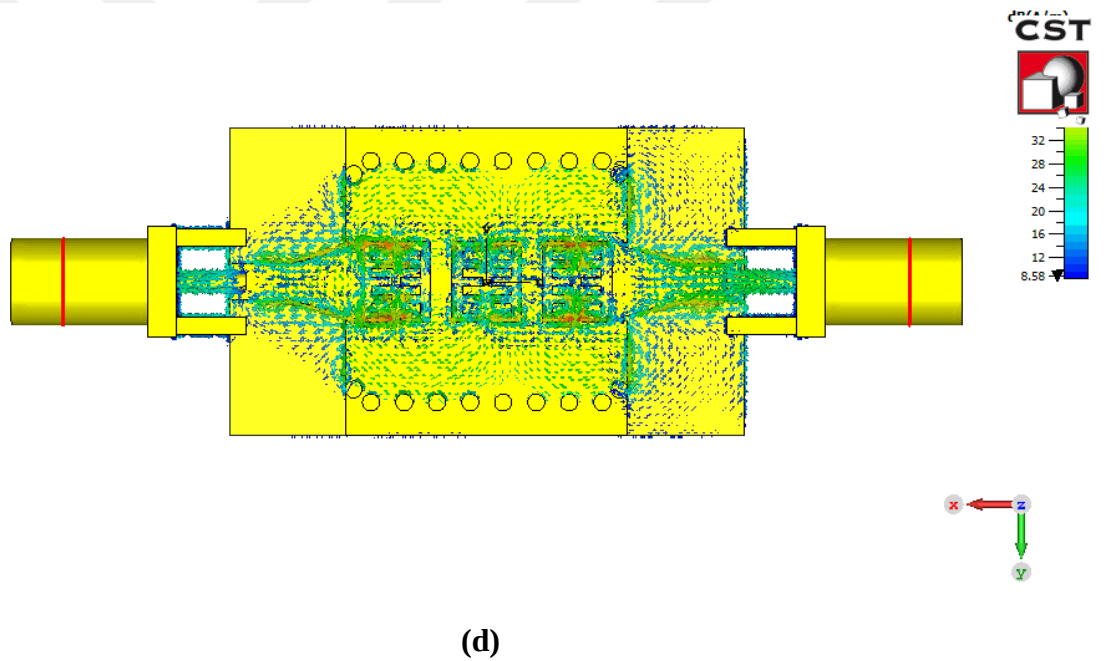
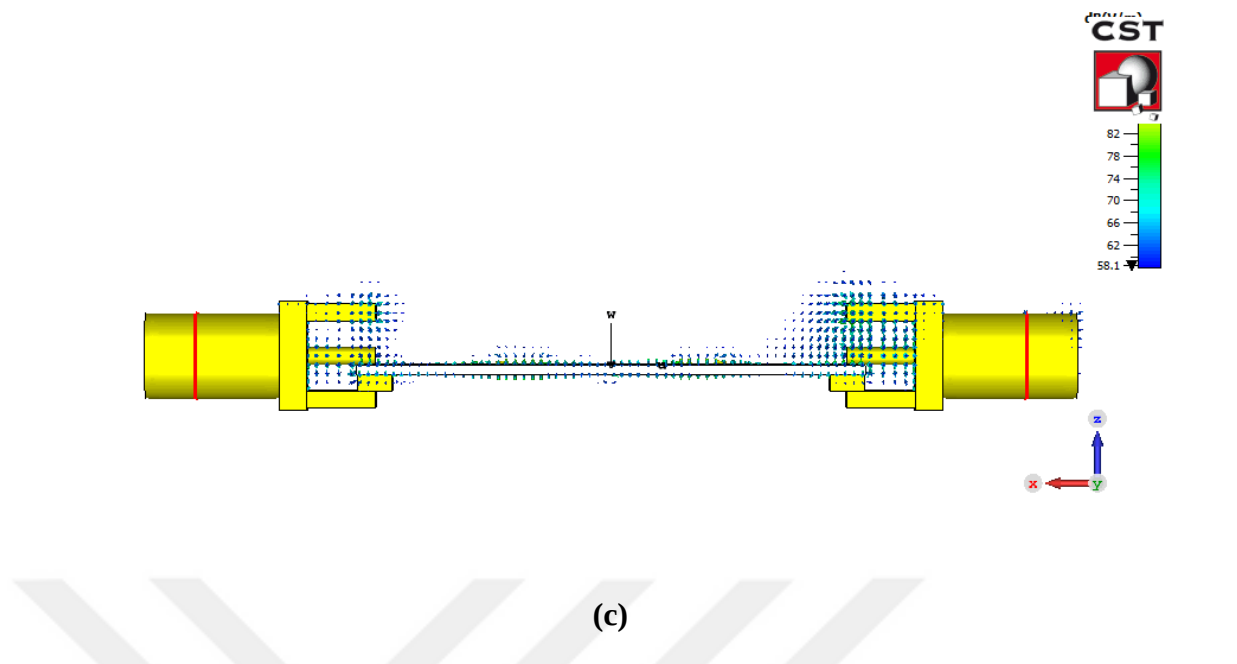


Figure 5. 7. Simulated S-parameters taking into account SMA bad installation vs. measured S-parameters

After looking at Figure 5.7, we conclude that these bad results are attributed to the wrong end installation of the SMA connectors used in measurement setup. Also, from the field distribution shown in Figure 5.8, it can be noticed that the inner conductor of the SMA that fall on the air act as external inductor disturb the matching between the end connectors and the filter's ports.





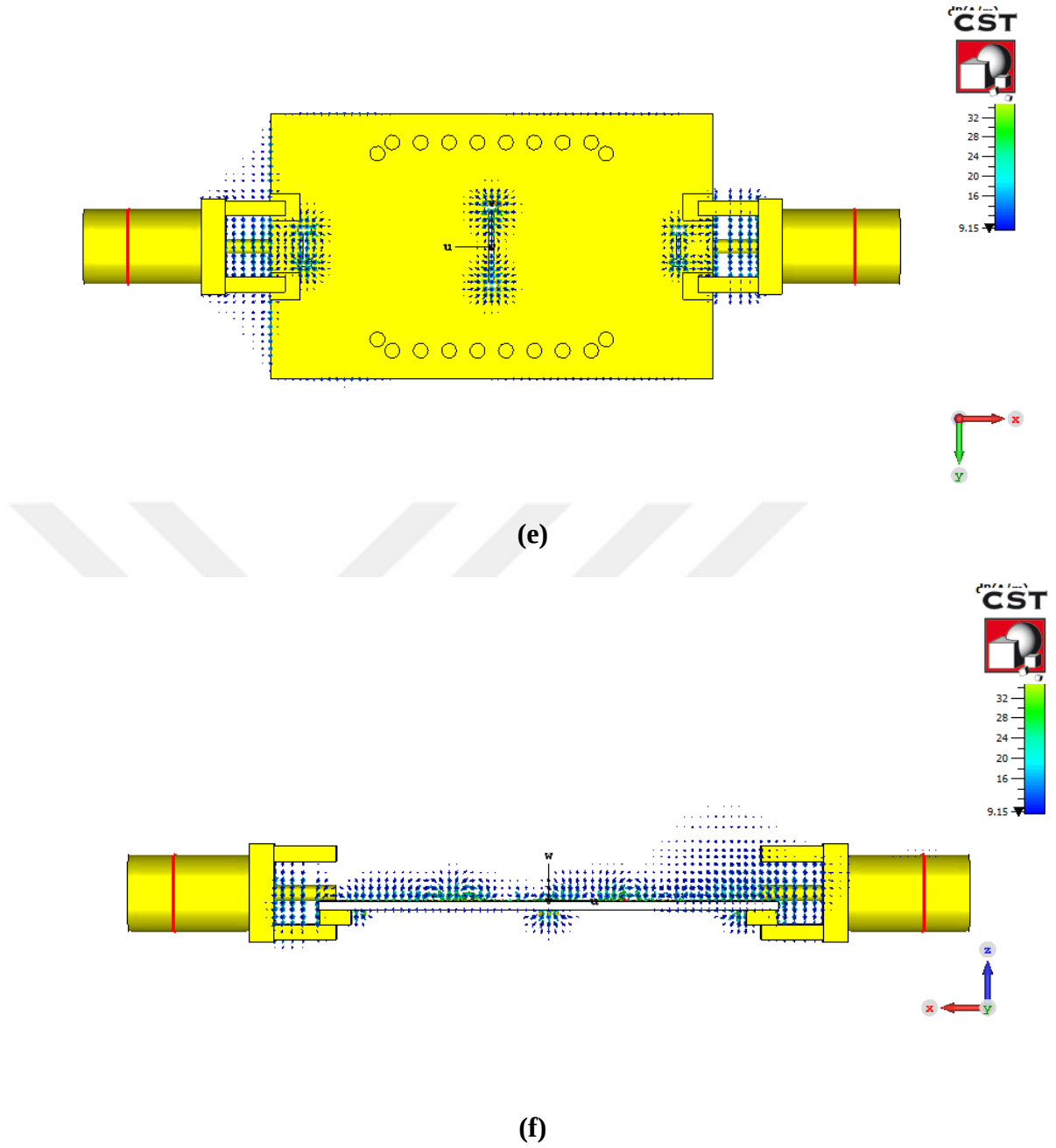


Figure 5. 8. Field distribution due to SMA miss-installation; **a)** E-field top view; **b)** E-field bottom view; **c)** E-field side view; **d)** H-field top view; **e)** H-field bottom view; **f)** H-field side view

So, while we were trying to avoid any problem could be originated due to any coupling between the SMA ground legs and dumbbells DGS we fell into destructive problem led to disturbing the matching between the filter and the connectors. Wherefore, we had corrected SMA installation and strengthen the soldering as it is shown in Figure 5.9, where now the SMA inner connectors are completely soldered with the ports, and the ground legs are soldered well to avoid any coupling affecting the measurements.

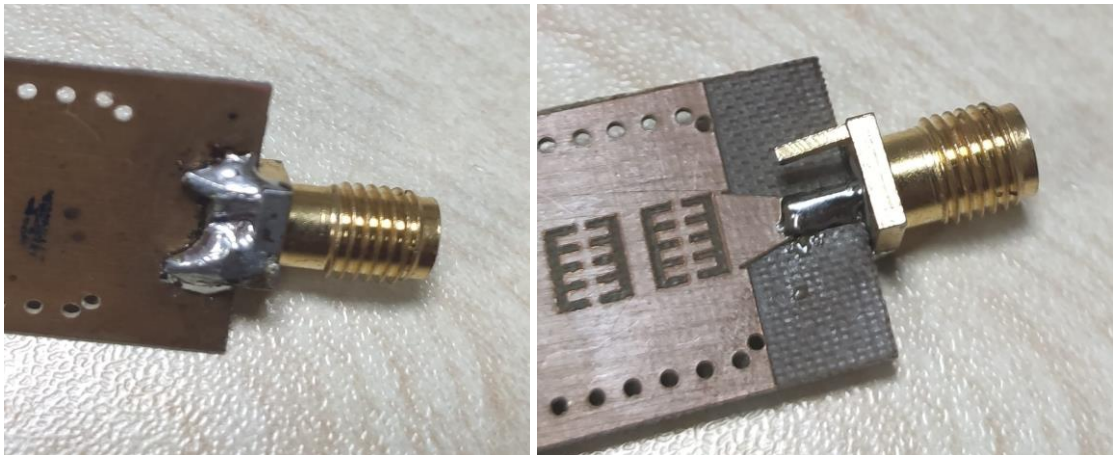
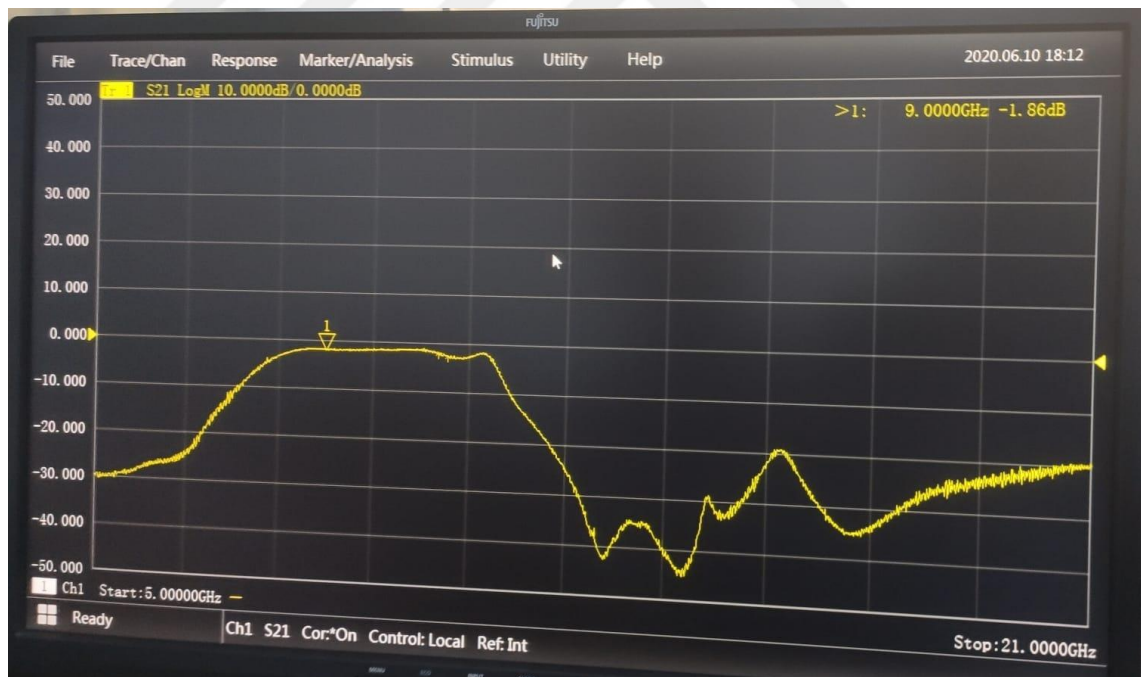


Figure 5. 9. Fixing/Correcting SMA connector installation

After the correction of SMA installation, we conducted the measurements and we got better results shown in Figure 5.10. comparison between the measured S-parameter response after correction SMA installation and the ideal simulated results is shown in Figure 5.11.



(a)



(b)

Figure 5. 10. Measured Results after the correction of SMA installation; **a)** S_{21} measured response; **b)** S_{11} measured response

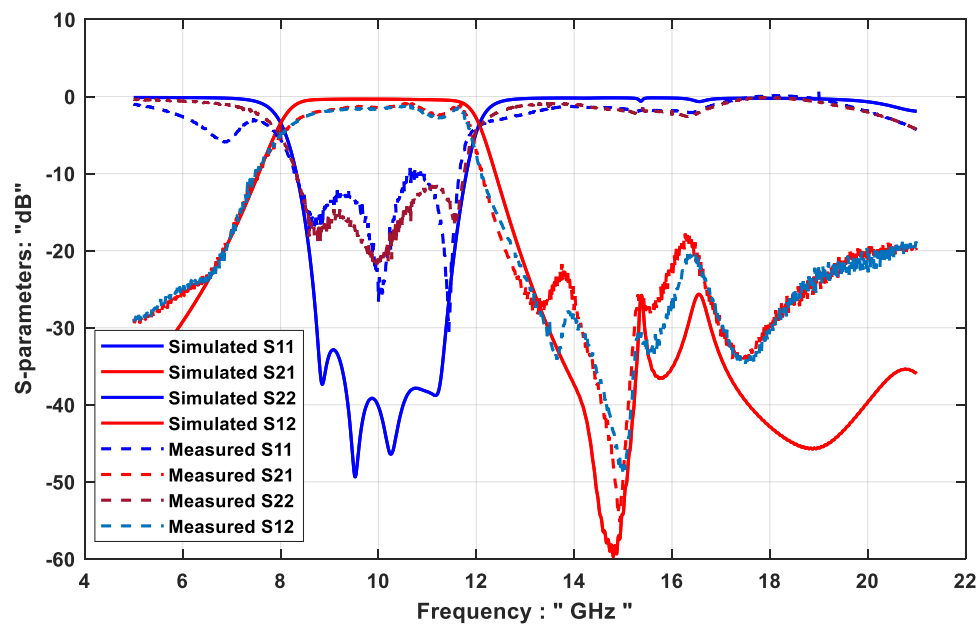
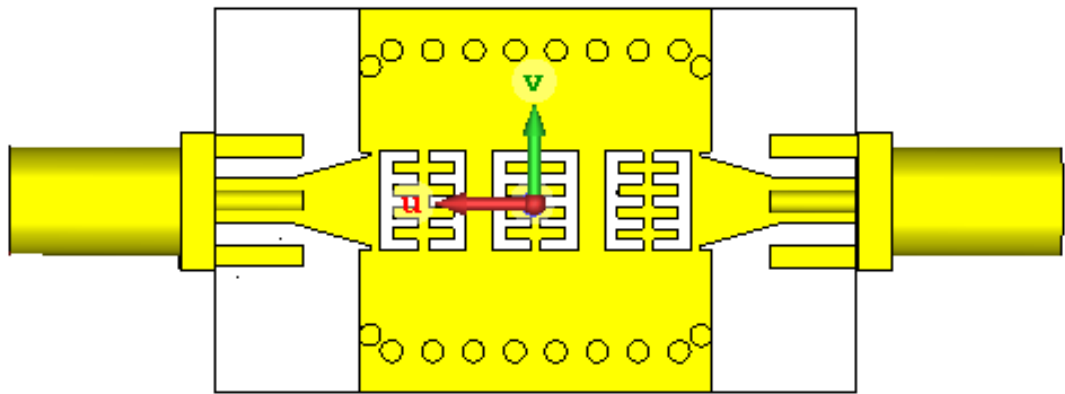
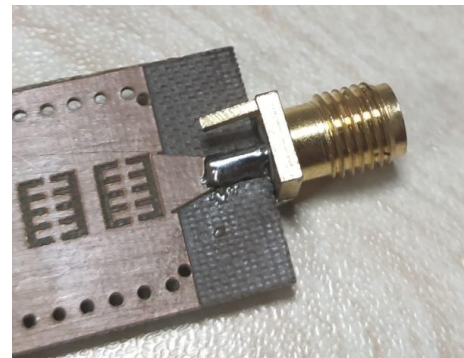
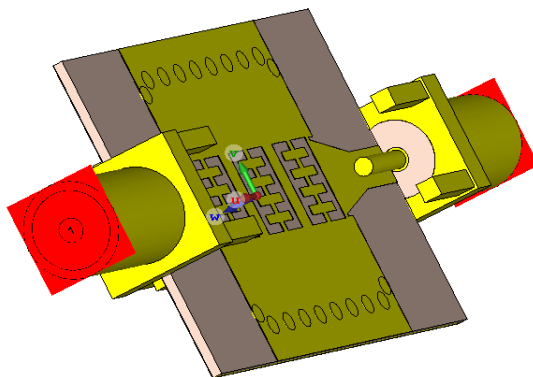


Figure 5. 11. Ideal simulation vs. measured S-parameters

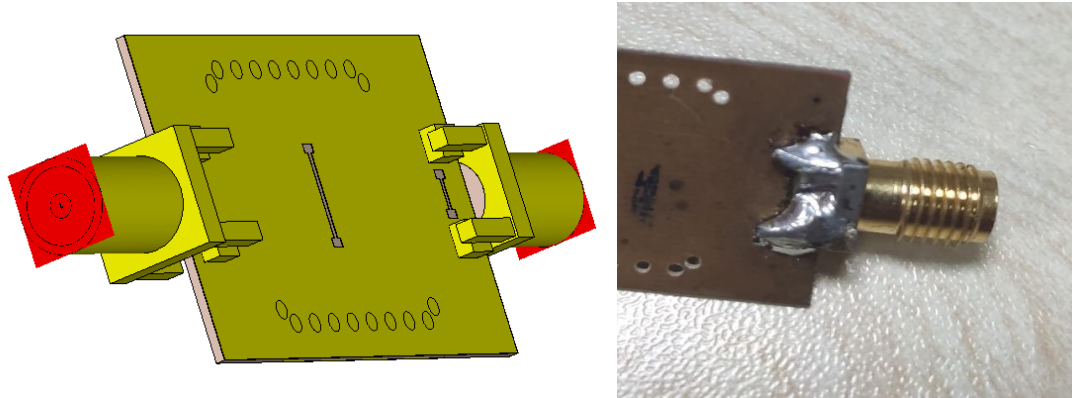
The measured transmission parameter (S_{21}) agree with the simulated (S_{21}) in term of center frequency and in term of out of band rejection, but there is insertion loss difference where the simulated insertion loss at the center frequency of 10 GHz about 0.33 dB while the measured response about 1.6 dB, also there is slight deviation at the transition's edges at the lower and the higher edges. In term of return loss as we see from Figure 5.11 there is a noticeable difference between to simulation and the measurements within operating band boundaries, so to introduce explanation to these issues we return to the simulation program to conduct the simulation taking into account the inclusion of the SMA connector almost exactly as in the measurements, SMA connector inclusion in the filter model is shown in Figure 5.12, a comparison between the modeling and the real installation of the connector is shown in Figure 5.12. In Figure 5.13 we present a comparison between the simulation results considering the SMA connector and the measurement response.



(a)



(b)



(c)

Figure 5. 12. Inclusion SMA connectors in the Simulator; **a)** top view; **b)** oblique view of the upper face; **c)** oblique view of the bottom face

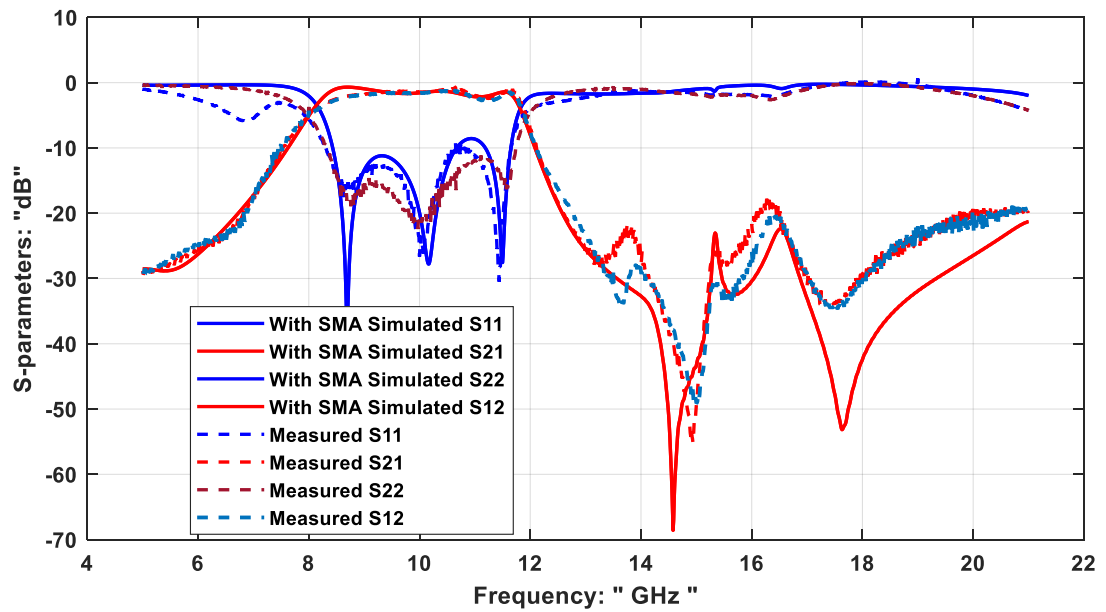


Figure 5. 13. Simulation SMA- included vs Measured results

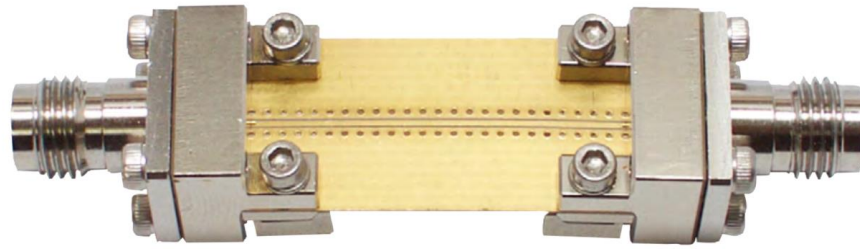
Here the measured results agree well with the simulation after the inclusion of the SMA in terms of the center frequency, bandwidth, return loss, and insertion loss as well as the out of band rejection, we have a measured insertion loss (S_{21}) about -1.6 dB, also, we have measured return loss (S_{11}) better than -10 dB, which agree well with simulation results that takes into account the SMA connector presence. at the lower edge frequency, we see that the transition of the measured response is less severe than the simulated this attributed to the transition fabrication not clear enough at the taper edges, and to the

drilling and coating quality of the via holes, that is because in these type of filters the via's dimensions and the distance between the vias affect the lower edge cut off frequency.

So, after looking at Figure 5.13, we conclude that the difference between the ideal simulation results and the measurements shown in Figure 5.11 attributed to end launch SMA connectors that are used in measurement setup but did not take into account during the design, these connectors affect the insertion and the return loss.

Also, high-performance end launch connectors that having a thinner and shorter inner pin connector, and of a separating distance between the inner conductor and ground legs suitable to laminate thickness should be used or using K-connector that fit well and work up to higher frequencies than SMA connectors, these connectors are shown in Figure 5.14, these type of connector provide excellent return Loss and better insertion losses up to frequencies more than 26 GHz.





(b)

Figure 5. 14. High-performance end launch connectors; **a)** high-performance SMA connector serve up to 26.5 GHz; **b)** 2.92mm or k- connector serves up to 50 GHz

5.2. Discussing the Reasons for the Results Variation

Here we present the main reasons that affect the agreement between the measurements and the simulation, then we suggest measures to get better compatibility between them.

The variation between the simulation and the measurement results is probably caused by:

- ☐ Fabrication tolerance.
- ☐ The connector losses and unexpected material losses that are not included in the simulations.
- ☐ Installation misalignment of the SMA connector, that is because the available connectors are not suitable for the filter, owing to SMA connector size does not suit the laminate size.
 - ❖ Should use a proper SMA of a thinner pin (inner conductor)
 - ❖ Using SMA connector having a separation distance between the inner pin (conductor) and the ground legs that match laminate's thickness used in the design process. The SMA connector we have used has a thick and a tall inner pin conductor also, the separation between the inner pin and the ground's legs is about 1.405 mm that is not matching and greater than laminate's thickness which is 0.508 mm. So, a high-performance connector with a tiny inner connector pin should be used, as suggested in section 5.1.

For better agreement between the simulation and the measurement:

- ☐ The design process and the real measurements should be done in a synchronized manner
 - ❖ To take into account any unexpected problems and differences that originated due to the material and conductor losses.
 - ❖ To ensure that set up of simulation settings gives reliable results.
- ☐ Inclusion the SMA during the design process to ensure solid taper transition tuning.

6. CONCLUSION

In this work, we have looked in the design processes of two bandpass filters based on Substrate integrated waveguide (SIW) and defected ground structures (DGS) techniques, that have the feature of being easily integrated with other planar structures. The first BPF is a wideband filter for X-band applications and the second is a broadband filter dedicated to C-band space communications. The modeling and simulation processes of the filters were conducted by building 3D models in CST Microwave Studio.

The X-band proposed filter gets a low insertion loss and good return loss with sharp and wide out-of-band rejection, we exhibited the design stages that includes the performance enhancement process. The filter prototype constructed using SIW-DGS cells in the upper face of the SIW and it had a good performance in the passband, but it suffers from spurious response appears in the stopband. To widen the bandwidth of the stopband and increase suppression of the spurious responses, dumbbell slots were etched on the ground plane of SIW resulting in a very wide stopband rejection and reduced spurious response, accompanied by better passband performance.

The C-band filter is designed for C-band satellite communication applications, it was mainly designed to simulate its peer filter of non-planar waveguide technology, which proved the ability of this cheap technology to do a good job when compared with the expensive bulky metallic waveguide's technology. Also, the proposed filters offer almost a flat group delay in the passbands. The proposed filters have a relatively compact size.

The outcomes of the work in this thesis spot the light on the validity and ability of this technology in designing high-performance wideband filters. Further size reduction can be done using Half Mode Substrate Integrated Waveguide. Other methods can be implemented along with SIW in designing of novel narrowband filters such as using coupling matrix synthesis to design filters based on iris type resonators or seeking for novel resonators with higher Q-factor, these filters can be supported with DGS or spur lines at the input and output ports to produce another transmission zeros to further enhance the stopband performance.

7. RESOURCES

- Ahn, D., Park, J., Kim, C., Kim, J., Qian, Y., & Itoh, T. 2001. A design of the low-pass filter using the novel microstrip defected ground structure. *IEEE Transactions on Microwave Theory and Techniques*, 49(1): 86-93.
- Caleffo, R. C. 2016. New Design Procedure to Determine the Taper Transition for Impedance Matching Between Microstrip Line and SIW Component. *Journal of Microwaves, Optoelectronics and Electromagnetic Applications*, 15: 247-260.
- Chen, X., & Wu, K. 2014. Substrate Integrated Waveguide Filter: Basic Design Rules and Fundamental Structure Features. *IEEE Microwave Magazine*, 15(5): 108-116.
- Deslandes, D., & Ke, W. 2006. Accurate modeling, wave mechanisms, and design considerations of a substrate integrated waveguide. *IEEE Transactions on Microwave Theory and Techniques*, 54(6): 2516-2526.
- Deslandes, D., & Wu, K. 2001. Integrated microstrip and rectangular waveguide in planar form. *IEEE Microwave and Wireless Components Letters*, 11(2): 68-70.
- Feng, X., & Ke, W. 2005. Guided-wave and leakage characteristics of substrate integrated waveguide. *IEEE Transactions on Microwave Theory and Techniques*, 53(1): 66-73.
- Hammerstad, E. O. 1975. Equations for microstrip circuit design, *Proceedings of the European Microwave Conference*: 268–272. Hamburg, Germany.
- Hirokawa, J., & Ando, M. 1998. Single-layer feed waveguide consisting of posts for plane TEM wave excitation in parallel plates. *IEEE Transactions on Antennas and Propagation*, 46(5): 625-630.
- HONG, J.-S. 2011. *Microstrip Filters for RF/Microwave Applications*. New York ;Chichester: Wiley.
- Jong-Im, P., Chul-Soo, K., Juno, K., Jun-Seok, P., Yongxi, Q., Dal, A., & Itoh, T. 1999. *Modeling of a photonic bandgap and its application for the low-pass filter design*. Paper presented at the 1999 Asia Pacific Microwave Conference. APMC'99. Microwaves Enter the 21st Century. Conference Proceedings (Cat. No.99TH8473).
- Ke, W., Deslandes, D., & Cassivi, Y. 2003. *The substrate integrated circuits - a new concept for high-frequency electronics and optoelectronics*. Paper presented at the 6th International Conference on Telecommunications in Modern Satellite, Cable and Broadcasting Service, 2003. TELSIKS 2003.
- Kordiboroujeni, Z., & Bornemann, J. 2014. New Wideband Transition From Microstrip Line to Substrate Integrated Waveguide. *IEEE Transactions on Microwave Theory and Techniques*, 62(12): 2983-2989.

- L. H. Weng, Y. C. G., X. W. Shi, and X. Q. Chen. 2008. AN OVERVIEW ON DEFECTED GROUND STRUCTURE. *Progress In Electromagnetics Research B*, 7: 173-189.
- Laboratory, G. L. R. a. M. I. o. T. R. 1964. *Microwave transmission circuits*: Boston Technical Publishers,.
- LANCASTER, J.-S. H. M. J. 2001. *Microstrip Filters for RF/Microwave Applications*.
- Liu, C. H., Xiang Hua. 2017. A SIW-DGS wideband bandpass filter with a sharp roll-off at upper stopband. *MICROWAVE AND OPTICAL TECHNOLOGY LETTERS*, 59.
- Oh, S., Koo, J., Hwang, M., Park, C., Jeong, Y., Lim, J., Choi, K., & Ahn, D. 2007. *An Unequal Wilkinson Power Divider with Variable Dividing Ratio*. Paper presented at the 2007 IEEE/MTT-S International Microwave Symposium.
- Pozar, D. M. 2012. *Microwave Engineering*. University of Massachusetts at Amherst: John Wiley & Sons, Inc.
- Sadiku, M. N. O. 2018. *Elements of Electromagnetics*: Oxford University Press.
- Steer, M. B. S. a. M. B. 2016. *FOUNDATIONS FOR MICROSTRIP CIRCUIT DESIGN*: John Wiley & Sons, Ltd.
- Tomassoni, C., Pelliccia, L., Cacciamani, F., Vallerotonda, P., Sorrentino, R., Galdeano, J., & Ernst, C. 2017. *Compact broadband waveguide filter with wide spurious-free range based on mixed TM and combline resonators*. Paper presented at the 2017 47th European Microwave Conference (EuMC).
- Wang, J., Hao, Z., & Kui-Kui, F. 2016. *A 110–150 GHz SIW-rectangular waveguide transition for terahertz applications*. Paper presented at the 2016 IEEE MTT-S International Microwave Workshop Series on Advanced Materials and Processes for RF and THz Applications (IMWS-AMP).
- Y. Cassivi, L. P., P. Arcioni, M. Bressan, K. Wu, and G. Conciauro. 2002. Dispersion Characteristics of Substrate Integrated Rectangular Waveguide *IEEE MICROWAVE AND WIRELESS COMPONENTS LETTERS*, 12.
- Y. Huang, Z. S., and L. Liu. 2013. A Substrate Integrated Waveguide Bandpass Filter Using Novel Defected Ground Structure Shape. *Progress in Electromagnetics Research-Pier*, 135: 201-213.
- Z. He, J. C., Z. Shao, X. Li, and Y. Huang. 2013. A Novel Power Divider Integrated with SIW and DGS Technology. *Progress In Electromagnetics Research*, 139: 289-301.

ÖZGEÇMİŞ

MOHAMMED R. A. NASSER

Mohammedr.alin@gmail.com



ÖĞRENİM BİLGİLERİ

Yüksek Lisans 2018-2020	Akdeniz Üniversitesi Mühendislik Fakültesi, Elektrik ve Elektronik Mühendisliği Bölümü, Antalya
Lisans 2012-2017	Gazze İslam Üniversitesi Mühendislik Fakültesi, Elektrik Mühendisliği Bölümü, Gazze, Filistin

MESLEKİ VE İDARİ GÖREVLER

Araştırma Görevlisi 2017-2018	Gazze İslam Üniversitesi Mühendislik Fakültesi, Elektrik Mühendisliği Bölümü, Gazze, Filistin
----------------------------------	---