Experimental Investigation of Microwave-Optoelectronic Interactions in a Microstrip Ring Resonator

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Abstract—**Microwave optoelectronic mixing is demonstrated on a semi-insulating gallium arsenide substrate, by monolithically integrating Schottky diode photodetectors into a microstrip ring resonator. When operated in the resistive mixing mode, a low frequency difference signal is extracted from the bias pad of the circuit. In the parametric mode, both degenerate and non-degenerate parametric amplification of an optical carrier signal takes place. The circuit shows good potential for application in wide-band fiberoptic systems.**

I. Introduction

Recent advances in optoelectronic technology have rendered multi-gigahertz bandwidth fiber-optic systems practical [1]. The excellent transmission properties of optical fibers and the immunity of lightwaves to electromagnetic interference, coupled with the availability of high speed laser diodes, have provided a major thrust to research in the field of microwave optoelectronics. This has made possible analog transmission of microwave signals over a fiber-optic cable by direct modulation of the laser diode [2]. In future microwave systems where signal and local oscillator waveforms are transmitted over optical fibers, analog microwave optoelectronic integrated circuits (OEIC's) with performance comparable to the present monolithic microwave integrated circuits (MMIC's) will be needed. Hence, integration of high frequency optical and electronic components in the same chip is currently under intense investigation [3].

The research reported herein pertains to a microwave OEIC in which a Schottky diode photodetector is integrated into a microstrip ring resonator. The flexibility afforded by this approach is that both the resistive and reactive changes of the photodetector, due to absorption of modulated light, can be independently availed to study microwave-optical interactions. When the study involves a resistive change, the circuit is said to be operated in the "resistive mixing mode." Studies associated with reactive changes involve operation of the circuit in the "parametric mode." The ensuing sections of this paper are devoted to the design and fabrication of the circuit and its characterization in the resistive mixing and parametric modes of operation.

Manuscript received March 28, 1991; revised July 24, 1991. This work was supported by the Office of Naval Research.

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IEEE Log Number 9103397.

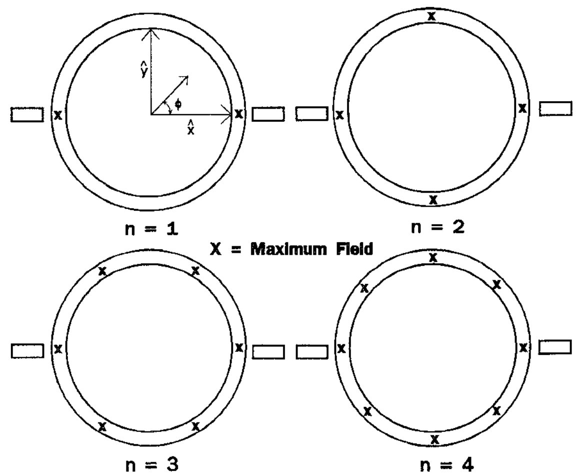

Fig. 1. Maximum electric field points for the first four modes.

II. Design

As the 3 dB modulation bandwidth of laser diodes gets pushed higher and higher in the microwave frequency spectrum, design of efficient optoelectronic components become very important. An important and essential component of an OEIC is the photodetector. To maximize transfer of signal energy from an optical source to a photodetector, the detector should be integrated into a circuit that resonates at the signal frequency; this stems from the fact that at resonance, energy is transferred from a source to a circuit at a maximum rate. At microwave frequencies, the two simple microstrip resonators are the linear and the ring resonator. The linear resonator radiates from both its open ends as opposed to the

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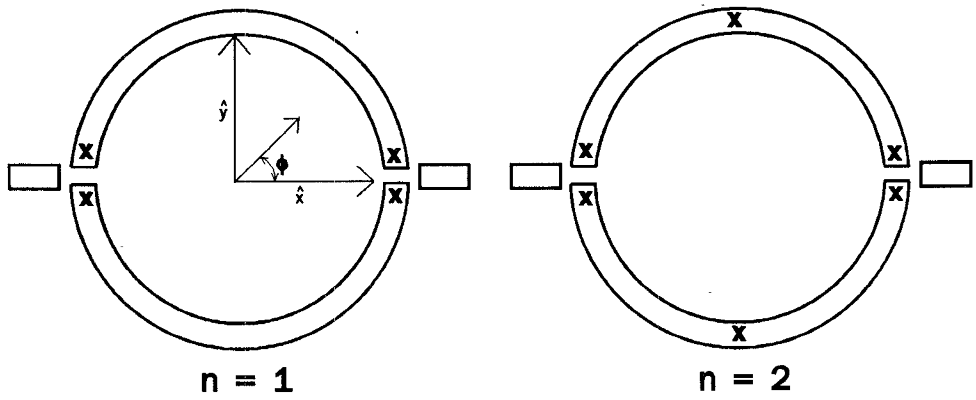


Fig. 2. Maximum electric field points of the first two modes for $\phi = 0°$ and 180°.

completely closed ring. Since radiation lowers the Q-factor of a circuit and hence the sharpness of resonance, coupling of microwave energy into a linear resonator is less efficient vis-a-vis the ring resonator. Hence, the ring was an obvious choice for the experiments reported in this paper.

To facilitate optical excitation of the ring resonator, slits have to be introduced to allow for integration of the photodetector. Concomitantly, this introduces the problem of field perturbation to be contended with. Fortunately, this problem can be alleviated by strategically locating these slits. The maximum electric field points for the first four modes of a completely closed ring are illustrated in Fig. 1. To minimize field perturbation, the slits have to be introduced such that field configuration of the resulting structure closely mimics that of Fig. 1. In the presence of slits, the fields in the resonator are altered so as to satisfy the corresponding boundary conditions in the resonator; these boundary conditions require that the electric field be at its maximum on either side of the slit. The fields of the ring resonator are least perturbed if the slits are located at azimuthal angles of $\phi = 0°$ and $\phi = 180°$ [4]. The field configuration for the first two modes of this structure is shown in Fig. 2 and these fields very closely mimic those of Fig. 1. Thus, by locating the photodetector at these slits, the intrinsic characteristics of the ring resonator are preserved. Also, when optical excitation is applied to one slit, the fields travelling in the clockwise and the counterclockwise directions add in phase at the other slit. Thus the ring supports standing waves, without excessive loss of optical power; this could be an efficient way to recover low power microwave signals transmitted over optical fibers. Also, this structure very closely mimics the characteristics of a completely closed ring because the radiated fields from the open ends of one half of the resonator, couples across the slits into the other half. Thus there is a closed-loop feedback of fields. Also, an independent microwave excitation can be applied to the resonator via feed lines skewed from the resonator so that there is little interaction between the microwave and optical signals. Hence, this configuration was applied to the resonator reported in this paper.

Design of resonators for optoelectronic applications is quite different from the design of conventional microwave resonators because integration of the photodetector must be considered. Two important design considerations applicable only in the optoelectronic domain are the spot size of the laser beam, and the transit time of the carriers; these factors dictate the choice of dimensions of the photodetector that should be integrated into the resonator. Large area photodetectors are more efficient collectors of carriers but generally have a poor frequency response because of the large carrier transit times. On the other hand, although small area detectors have a better frequency response, it is considerably more difficult to couple light into the detector. Photodetectors that respond at microwave frequencies typically have electrode spacings of the order of a few microns; these electrodes are biased so that the carriers traverse at the saturation velocities. For GaAs, the saturation velocity is approximately 10^7 cm/s. In GaAs, an electron travelling at the saturation velocity takes 40 ps to traverse a 4 μm gap; this time is much smaller than the speed of response needed to track a 10 GHz signal, which corresponds to 100 ps. In view of the process tolerances that could be afforded and to enable efficient focusing of light, an electrode spacing of 4 μm was chosen. Bias pads were designed in conjunction with the resonator to enable application of bias to the photodetector; this again was done with minimal field perturbation to the resonator. The lowpass filters that constitute the bias pads were designed so that their passband encompassed the desired range of baseband frequencies (a few hundred MHz for this research), for microwave optoelectronic mixing. Thus by placing a bias-T at the bias pad, difference frequencies at baseband can be extracted at the bias port; sum and difference frequencies in the microwave band can be extracted at one of the feed lines to the resonator.

Illustrated in Fig. 3 is the layout of the microwave optoelectronic ring resonator circuit. The connections to

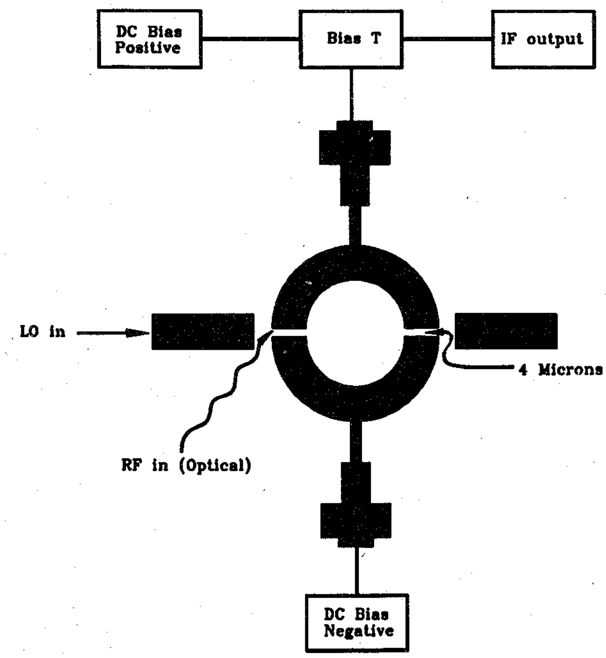


Fig. 3. Layout of the ring resonator circuit.

the different ports of this figure correspond to experiments in the resistive mixing mode which will be discussed later. The ring is designed to operate at a fundamental frequency of approximately 3.5 GHz and has a mean radius of 4.73 mm. Two 4 μm slits are introduced at diametrically opposite locations of the ring for optical excitation. A dc bias across the gaps is applied via low-pass filters with a cutoff frequency near 2 GHz. Since this frequency is lower than the fundamental resonant frequency of the ring, the intrinsic fields of the resonator are not appreciably perturbed [4]. The dimensions of coupling gaps between the feed lines and the resonator were chosen to be 30 μm and 100 μm, respectively and the microwave local oscillator (LO) excitation is applied via the more loosely coupled 100 μm gap; the output beat signal is extracted across the 30 μm gap. It is thus ensured that whereas the LO signal is loosely coupled into the resonator, extraction of the output signal is more efficient due to the tighter coupling associated with the 30 μm gap.

III. Fabrication, Operation and Characterization

The circuit is fabricated on semi-insulating GaAs substrate by photolithograhically delineating the pattern on a 1000 Å thick Au-Ge-Ni layer. This is followed by electroplating with Au to a thickness of about 3.2 μm. The Ni and Au-Ge are then removed from the unpatterned areas by etching and the substrate is annealed in a resistive strip heater at 240°C.

Under application of bias, each slit acts like a pair of diodes connected back to back; whereas just one slit is used for optoelectronic mixing, the other is needed to bias the circuit. Current conduction in these devices may be explained as follows: When the back to back pair of diodes is biased at a certain critical field, breakdown occurs. The carriers traverse the 4 μm gap region and get collected at the other contact. In the resistive mixing mode of operation of the ring resonator, the detector diode is biased to operate in the breakdown region. As long as the maximum junction temperature is not exceeded, the device is not destroyed due to operation in the breakdown region.

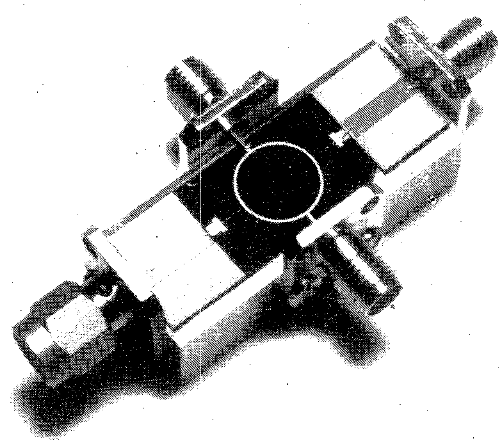


Fig. 4. Illustration of the packaged mixer circuit.

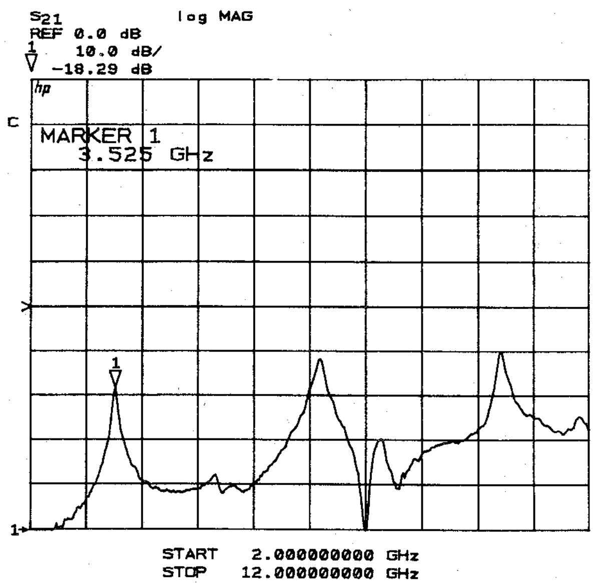


Fig. 5. $|S_{21}|$ versus frequency of the circuit.

The circuit is glued down to a brass block by a conductive silver epoxy. Also glued to the brass block are two RT/Duroid strips that are bonded to the bias pads. SMA connectors are attached to the RF and bias ports of the circuit which is pictorially illustrated in Fig. 4. The RF characteristics of the device was measured using the HP 8510B automatic network analyzer. The first five resonances occurred at 3.52, 7.18, 10.4, 13.52 and 16.9 GHz, respectively; the corresponding loaded Q-factors are 44,

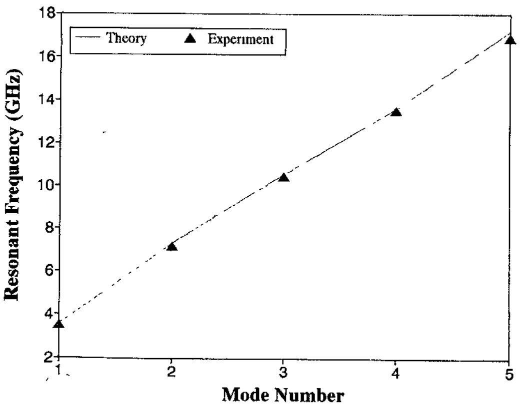


Fig. 6. Mode number versus frequency; comparison of theory and experiment.

59.83, 74.29, 56.33 and 105.62. The $|S_{21}|$ vs frequency characteristics of the first three modes are shown in Fig. 5. The circuit was modeled using distributed transmission line theory in which the ring was modeled as a 36-sided polygon. The discontinuities at the vertices of the polygon and the gaps between the feed line and the resonator were modeled by their appropriate equivalent circuits [5]; the slits were modeled as shorts, and the bias pads were modeled as open transmission line stubs. The theoretical and experimental results for resonant frequency are compared in Fig. 6 and the agreement is quite good.

IV. Experiments in the Resistive Mixing Mode

In a resistive mixer, a weak radio frequency (RF) signal, $V_{RF}\cos(\omega_{RF}t)$, is mixed in a time-varying resistance, with a much stronger local oscillator (LO) signal $V_{LO}\cos(\omega_{LO}t)$. The resulting current contains Fourier components of frequencies $n\omega_{LO} \pm m\omega_{RF}$, where n and m are integers. However since $V_{RF} \ll V_{LO}$, the only frequency components that are of significance are $n\omega_{LO} \pm \omega_{RF}$. The intermediate frequency (IF) is defined as

$$\omega_{IF} = |\omega_{LO} - \omega_{RF}| \quad (1)$$

where ω_{RF} can be either larger or smaller than ω_{LO}. The frequency components related to the RF signal can be expressed as

$$n\omega_{LO} \pm \omega_{RF} \equiv n'\omega_{LO} \pm \omega_{IF} \quad (2)$$

where $n = 1, 2, 3, \cdots$, and $n' = 0, 1, 2, \cdots$. The frequencies $n'\omega_{LO} \pm \omega_{IF}$ are referred to as the harmonic sidebands; the signal voltage will cause current to flow at all these frequencies indicating that it is possible to transfer signal power to these frequencies. When the circuit is operated in the resistive mixing mode, the RF signal pumped into the resonator is a directly modulated optical signal from a laser diode; this signal is mixed with a much larger LO signal from an independent microwave source and the difference signal is obtained at the bias pad. In this case, the photodetector in the slit region also serves as the mixing element due to the time-varying nature of its conductance.

Illustrated in Fig. 3 are the port connections to the ring resonator in the resistive mixing mode. DC bias is applied between the two bias pads. The Bias-T provides the necessary isolation between the dc and the IF signal paths. The low-pass filters designed in conjunction with the bias pads have a cutoff frequency of approximately 2 GHz and hence block the fundamental resonance of the ring. However, the IF frequency that is of the order of a few hundred MHz is effectively passed by the filter. Since the ring resonator naturally acts like a bandpass filter, the RF and LO frequencies have to be close to one of the ring's resonances to be able to observe mixing of optical and microwave signals.

The block diagram of the experimental setup is illustrated in Fig. 7. The laser diode, collimating lenses and the device under test are mounted on *X-Y-Z* micropositioners secured to an optical table. The Ortel SL 1010 laser diode (lasing wavelength = 0.84 μm, threshold current is 6.6 mA) is biased at 9 mA, and is directly modulated by a HP 8340A microwave synthesized sweeper. The input microwave power level into the laser diode is typically around -19 dBm. The output of the device under test was fed into a HP 8565A spectrum analyzer. To facilitate mixing of microwave and optical signals, microwave excitation (LO) was independently applied by means of a HP 8609B sweep oscillator; this excitation is applied via the feed line skewed from the resonator by 100 μm. To characterize the photodetector, light was shone in the slit and dc current through the slit region was monitored as a function of voltage. The bias dependence of dark and photocurrents across the slits are shown in Fig. 8. When the bias voltage is 8 V or greater, dark current approaches photocurrent. This is because there is a large concentration of carriers from the background dark current due to application to bias, and hence further enhancement of current due to photoinjection is very small. Current detection even at zero bias is possible due to a photovoltaic effect that modifies the built-in potential of the junction, thereby providing a small effective forward bias.

When a modulated optical signal (RF) is applied to one of the slits of the ring resonator, an RF voltage is induced; this effect is enhanced if the frequency of the signal is near one of the ring's resonances. If an independent microwave signal (LO) also at the same resonance is applied via one of the feed lines, then the difference frequency (IF) can be obtained at the bias pad. To be able to efficiently extract the difference frequency (IF) at the bias pad, this signal has to be less than 2 GHz, which is the cutoff frequency of the low-pass filter. If the RF and LO are at different resonances of the ring, then the IF would be larger than 3 GHz, and hence cannot be detected at the bias pad. The experiments reported in this section correspond to operation of the circuit with both LO and RF in the vicinity of the ring's first resonance.

When a modulated optical signal at 3.467 GHz is applied to one of the slits, and a LO at 3.596 GHz is applied at the feed line, the spectrum of IF signal obtained at the

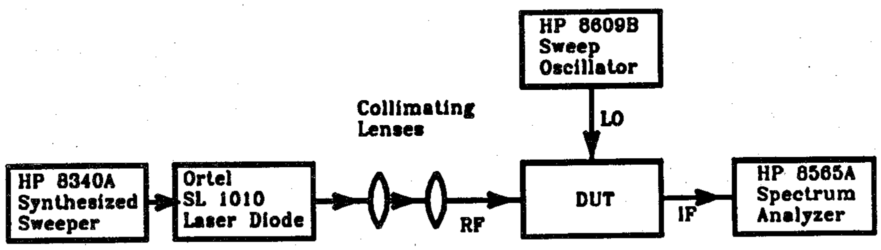


Fig. 7. Experimental test setup.

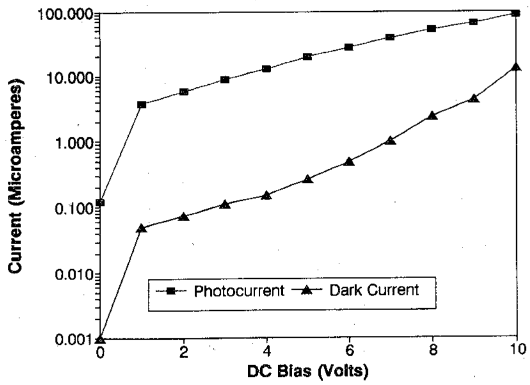


Fig. 8. Dark and photocurrent across the slits versus bias voltage.

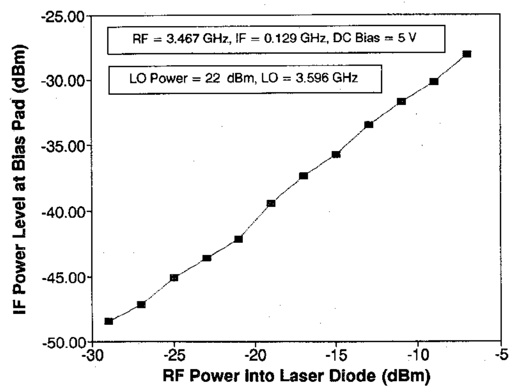


Fig. 10. IF power at bias pad versus RF power into laser diode.

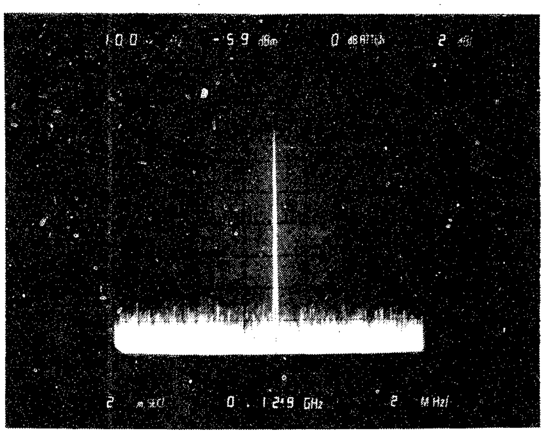


Fig. 9. Spectrum of IF output at bias pad. RF = 3.467 GHz, LO = 3.596 GHz, IF = 0.129 GHz.

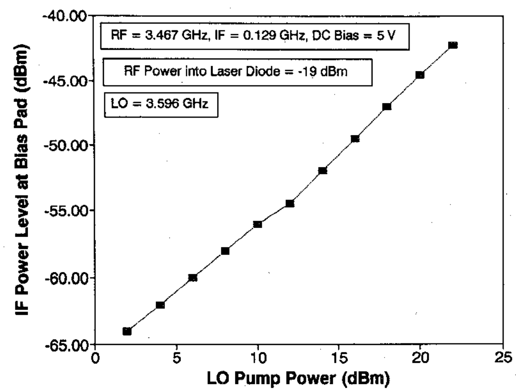


Fig. 11. IF power at bias pad versus LO pump power.

bias pad is shown in Fig. 9. The signal increases linearly with bias before reaching saturation [6]. However, this figure corresponds to fixed RF and LO pump powers. The dependence of the IF signal on both the RF and LO microwave pump powers are shown in Figs. 10 and 11, respectively. As expected, the signal increases linearly in both cases. The advantages of using the ring resonator circuit in this configuration are many. First of all, the LO and RF signals are mutually isolated. The RF signal being optical, is immune to electromagnetic interference. Hence the microwave LO port and the optical RF port are mutually decoupled. Thus the LO to RF isolation is almost infinite. On the other hand, since the feed line that pumps LO is 100 μm away from the resonator, very little RF can leak into the LO port. In all cases, the RF to

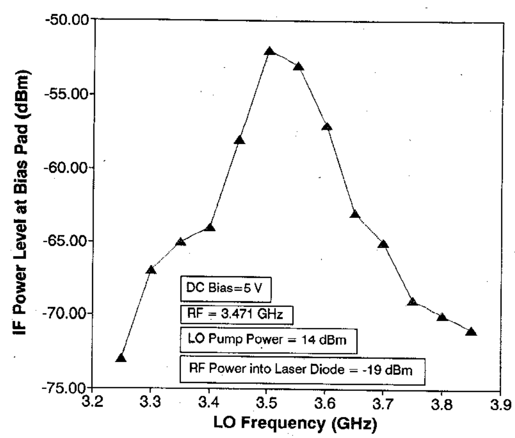


Fig. 12. IF power at bias pad versus LO frequency.

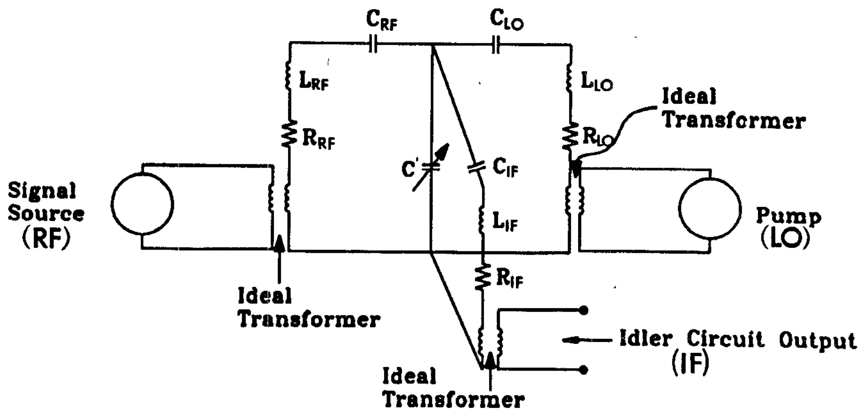


Fig. 13. Equivalent circuit of a parametric amplifier.

LO isolation that can be obtained in this configuration is far superior to what can be expected in conventional microwave mixers. Another advantage of this configuration is that the image frequency which is given by $2f_{LO} - f_{RF}$ is automatically suppressed due to the narrow passband at resonance. Since this frequency lies just outside of ring's first resonance, it is not passed by the low-pass filter. Also, since the ring does not support this frequency, the image frequency cannot be extracted from the feed line either. As mentioned earlier, to be able to observe resistive mixing, the RF and LO frequencies have to be in the vicinity of the ring's resonance. If either one of these frequencies is slowly tuned away from resonance, the IF signal strength at the bias pad gradually decreases. This is illustrated in Fig. 12. As can be seen, the peak of the IF signal output occurs when the LO is close to the ring's resonance; when it is tuned out of resonance the strength of the IF signal goes down. Although the LO was varied in this case, similar effects were observed in varying the RF.

V. Experiments in the Parametric Mode

A. Nondegenerate Parametric Amplification

The basic difference between the resistive mixing mode and the parametric mode is that, in the parametric mode, the nonlinear element is a time-varying reactance; in the case of the ring resonator, this time-varying reactance is the capacitance of the slit region. The other factors unique to the parametric mode are that there is no dc bias required, and the output beat signal is extracted from the feed line. In the case of the ring resonator, parametric effects are observed when the RF, LO and IF are at a ring's resonance. Illustrated in Fig. 13 is the equivalent circuit of a parametric amplifier. R, L and C correspond to the resistance, inductance and capacitance of the circuit with subscripts RF, LO and IF denoting the RF, LO and IF circuits, respectively. C' is the time-varying nonlinear capacitance. The circuit has three resonant frequencies Ω_{RF}, Ω_{LO} and Ω_{IF} corresponding to the signal, pump and idler circuits, respectively. These frequencies are given by

$$\Omega_{RF} = \frac{1}{\sqrt{L_{RF}C_{RF}}}$$
$$\Omega_{LO} = \frac{1}{\sqrt{L_{LO}C_{LO}}}$$
$$\Omega_{IF} = \frac{1}{\sqrt{L_{IF}C_{IF}}} \qquad (3)$$

In the subsequent discussions, it will be assumed that the frequencies ω_{RF}, ω_{LO} and ω_{IF} are equal to the resonant frequencies Ω_{RF}, Ω_{LO} and Ω_{IF}, respectively.

If a load resistance R_L is connected across the output terminals of the idler circuit output, and if L_{IF} and C_{IF} are adjusted so that ω_{IF} is equal to $\omega_{LO} - \omega_{RF}$, appreciable power develops across R_L. If the LO power is much stronger than the RF power, the voltage across R_L will be much larger than the RF signal voltage. Most of the load power thus comes from the pump source. If $\omega_{IF} < \omega_{RF}$, the circuit is known as a *parametric downconverter*. On the other hand, if $\omega_{IF} > \omega_{RF}$, the circuit is known as a *parametric upconverter*. In either case, the signal at one frequency is converted to an amplified output at a different frequency. However, if $\Omega_{LO} = 2\Omega_{RF}$, then $\Omega_{IF} = \Omega_{RF}$, and amplification takes place at the RF signal frequency. In this case, the circuit is said to be a *degenerate parametric amplifier*.

In the parametric mode experiments, the LO pump signal is fed into the resonator from the feed line across the 100 μm gap, and the output IF signal is extracted across the 30 μm gap at the other feed line. When a LO pump signal at 3.52 GHz (pump power = 22 dBm) is applied to the feed line, and a 3.537 GHz RF optical signal is applied to the slit, a sum frequency of 7.057 GHz is obtained at the output; this sum frequency corresponds to the ring's second resonance. When the LO pump frequency is at 7.07 GHz (ring's second resonance), and the RF signal frequency at 3.485 (ring's first resonance), a sum frequency at 10.557 GHz (ring's third resonance) is obtained. The spectrum of the sum frequencies at the ring's second and third resonances are shown in Figs. 14

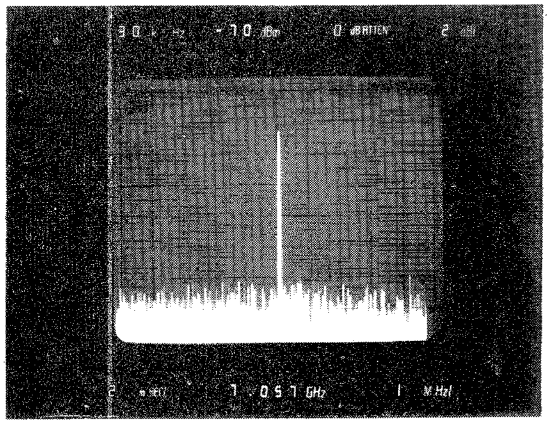


Fig. 14. Spectrum of IF output at feed line. RF = 3.52 GHz, LO = 3.537 GHz, IF = 7.057 GHz. LO pump power = 22 dBm, RF power into laser = − 19 dBm.

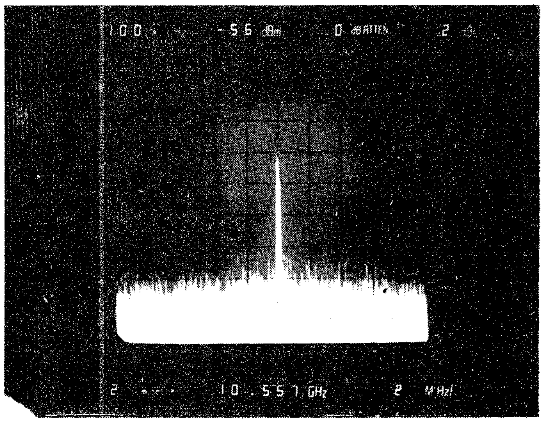
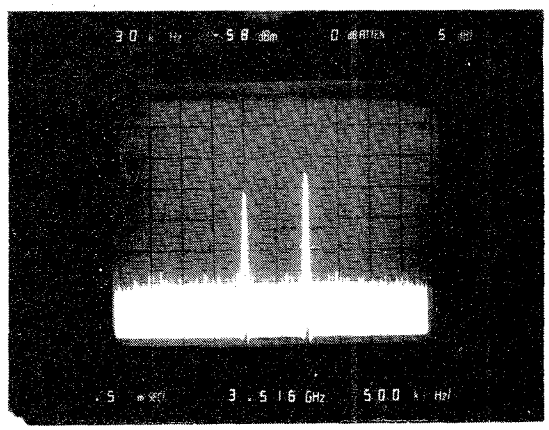


Fig. 15. Spectrum of IF output at feed line. RF = 3.487 GHz, LO = 7.07 GHz, IF = 10.557 GHz. LO pump power = 22 dBm, RF power into laser = − 19 dBm.

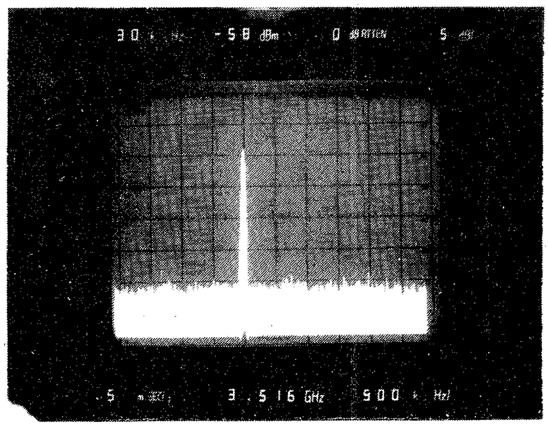
and 15, respectively. When the LO pump and RF signal frequencies are at the ring's first resonance, the difference frequency (IF) can be extracted at the bias pad in the resistive mode. This was discussed in the previous section. However, when the pump frequency (LO) is at the ring's second resonance, and the optical signal (RF) at the first, the difference frequency (IF) will be equal to RF. Operation of the circuit in this mode, will be explained in the next section dealing with degenerate parametric amplification.

B. Degenerate Parametric Amplification

Degenerate parametric amplification occurs when $\omega_{LO} = 2\omega_{RF}$ [7]. When this occurs, since $\omega_{IF} = \omega_{RF}$, amplification takes place at the RF signal frequency. This is clearly illustrated in Fig. 16. The optical RF signal is fixed near the ring's first resonance and the pump LO signal is varied in the vicinity of the ring's second resonance. Shown in Fig. 16(a) is the spectrum of the IF and RF signals just before the variation of LO causes the RF and IF signals to overlap; the signal on the left is the RF signal, and the signal on the right is the IF signal. As can be seen, the IF power level in the spectrum is greater than the RF power level, indicating that there is already some amplification. This effect is further enhanced when the LO is varied to cause the RF and IF frequencies to overlap. This is shown in Fig. 16(b). A 7 dB enhancement is observed when the RF and IF overlap. Shown in Fig. 17 is the dependence of IF power on the LO pump power. The dependence as expected, is almost linear.



(a)



(b)

Fig. 16. Spectrum of IF output at feed line. (a) Before overlap. (b) After overlap of RF and IF. LO pump power = 22 dBm, RF power into laser = − 19 dBm.

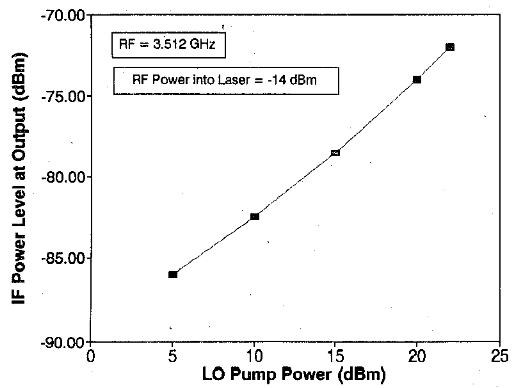


Fig. 17. IF power versus LO pump power for degenerate parametric amplification.

Before concluding, it must be re-emphasized, that for operation in the parametric mode, the ring resonates at the RF, LO, and IF frequencies. Since the ring has multiple resonant frequencies, the circuit can be operated only if these three frequencies happen to be near one of the ring's resonances. Hence, the bandwidth of the circuit is very limited. However, by integrating a varactor into the ring, the resonant frequencies can be electronically tuned [8], [9] to enhance the mixing bandwidth.

VI. Conclusion

In conclusion, we have experimentally investigated the interaction of microwave and optical signals in a microstrip ring resonator on semi-insulating GaAs substrate. When the signals are near resonances of the ring resonator, the interaction is enhanced. This type of optoelectronic resonator circuit can be used for generating IF signals in receivers for analog fiber optic communication and in OEIC's for optically controlled microwave oscillators.

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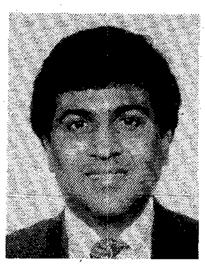
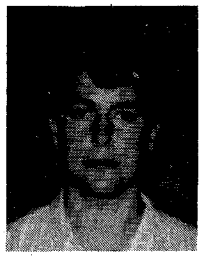


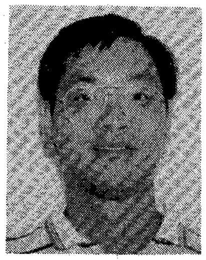
Ganesh K. Gopalakrishnan (S'85–S'88–M'88–M'89–S'90–M'90) received the B.S. degree from the University of Madras, India, in 1983, the M.S. degree from the University of Oklahoma Norman, in 1985 and the Ph.D. degree from Texas A&M University, College Station, in 1990 in electrical engineering. At Texas A&M he conducted research in the fields of microwave circuits and microwave-optical interactions and has two patent applications.

He is now with the Maryland Advanced Development Laboratory at the Optical Sciences Division of the Naval Research Laboratory in Washington DC. His present research interests include high-speed integrated optical devices and microwave-optical interactions.



Brock W. Fairchild was born in Houston, TX on December 29, 1965. He received the B.S. and M.S. degrees in electrical engineering from Texas A&M University, College Station, in 1988 and 1990, respectively.

From 1989 to 1990 he worked as a Research Assistant at the Solid State Electronics Laboratory at Texas A&M University. In the summer of 1989 he worked at Advanced Micro Devices in the Diffusion Group. He is currently employed at Texas Instruments, Dallas, TX, as a Process Engineer in the 0.8 Micron BICMOS Process Development Group.

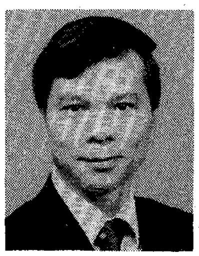


Chun-Liang Yeh was born in Cha-Yi, Taiwan, Republic of China in 1963. He received B.S. degree in physics from Tsinghua University, Taiwan in 1985. He is currently working toward the M.S.E.E. degree at Texas A&M University, College Station, TX.

His present research involves process development for MMIC devices.

Chang-Soo Park was born in Seoul, Korea on January 1, 1955. He received the B.S. degree in electronic engineering from Hanyang University in 1979 and the M.S. degree in electronic engineering at Seoul National University in 1981.

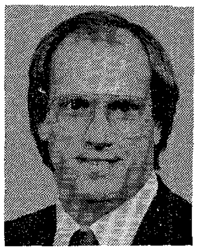
After graduation, he joined the Technical Staff in the Electronics and Telecommunications Research Institute (ETRI) and was a Senior Member of the Technical Staff before he came to Texas A&M University in September 1986. After receiving the Ph.D. Degree in Electrical Engineering at Texas A&M in December 1990, he returned to ETRI.



Kai Chang (S'75–M'76–SM'85–F'91) received the B.S.E.E. degree from the National Taiwan University, Taipei, Taiwan, the M.S. degree from the State University of New York at Stony Brook, and the Ph.D. degree from the University of Michigan, Ann Arbor, in 1970, 1972, and 1976, respectively.

From 1972 to 1976, he worked for the Microwave Solid-State Circuits Group, Cooley Electronics Laboratory of the University of Michigan as a Research Assistant. From 1976 to 1978, he was employed by Shared Applications, Inc., Ann Arbor, where he worked in computer simulation of microwave circuits and microwave tubes. From 1978 to 1981, he worked for the Electron Dynamics Division, Hughes Aircraft Company, Torrance, CA., where he was involved in the research and development of millimeter-wave solid-state devices and circuits, power combiners, oscillators and transmitters. From 1981 to 1985, he worked for the TRW Electronics and Defense, Redondo Beach, CA., as a Section Head, developing state-of-the-art millimeter-wave integrated circuits and subsystems including mixers, VCO's, transmitters, amplifiers, modulators, upconverters, switches, multipliers, receivers, and transceivers. He joined the Electrical Engineering Department of Texas A&M University in August, 1985 as an Associate Professor and was promoted to a Professor in 1988. In January 1990, he was appointed E-Systems Endowed Professor of Electrical Engineering. His current interests are in microwave and millimeter-wave devices and circuits, microwave integrated circuits, microwave optical interactions, and antennas.

He served as the Editor of the four-volume *Handbook of Microwave and Optical Components* (Wiley, 1989, 1990). He is the Editor of the *Microwave and Optical Technology Letters* and the *Wiley Book Series in Microwave and Optical Engineering*. He has published over 130 technical papers and several book chapters in the areas of microwave and millimeter-wave devices and circuits. Dr. Chang received the Special Achievement Award from TRW in 1984, the Halliburton Professor Award in 1988 and the Distinguished Teaching Award in 1989 from the Texas A&M University.



Mark H. Weichold (S'75–S'78–S'80–S'81–M'81–M'82–SM'90) received the B.S., M.S., and Ph.D. degrees in electrical engineering from Texas A&M University, College Station.

He joined the Electrical Engineering faculty at Texas A&M University in September 1982, where he presently holds the rank of Associate Professor. He has also worked for General Dynamics Ft. Worth Division, Motorola in Austin, and the U.S. Army Electronics Technology and Devices Laboratory in Ft. Monmouth, NJ. He is active in consulting with companies such as MCC, Motorola, Los Alamos National Labs, Electro Scientific Industries, US Intec, and Dallas Semiconductor. He specializes in electron device research including process development, device design, fabrication, and characterization. Current research activities include investigations in vacuum microelectronics, solid state nanoelectronics, novel lithographic techniques, optical microwave devices, and IR spatial modulators. Other activities include fabrication of micro heat pipes, ion beam lithography and detector development, IC manufacturing research, and molecular beam epitaxy.

Dr. Weichold is a Registered Professional Engineer in the State of Texas. He has authored over 35 journal articles, conference papers, and significant reports and has four patent applications pending. He has also chaired several IEEE conference sessions and is a Senior Member of the American Physical Society and the American Vacuum Society. He has been awarded the Eta Kappa Nu Outstanding Professor Award in 1987 and 1991, and the General Dynamics Excellence in Engineering Teaching Award in 1989.

Henry F. Taylor (SM'78–F'85) was born in Ft. Worth, Texas, on September 27, 1940. He received the B.A., M.A., and Ph.D. degrees in physics from Rice University in 1962, 1965, and 1967, respectively.

He was employed as a Research Physicist at the Naval Ocean Systems Center (formerly the Naval Electronics Laboratory Center) in San Diego, CA from 1967 to 1978. From 1978 to 1980 he was employed by Rockwell International in Thousand Oaks, CA, where he was Principal Scientist of the Optoelectronics Department of the Microelectronics Research and Development Center. From 1980 to 1985 he was Head of the Optical Techniques Branch of the Naval Research Laboratory in Washington, DC. He joined the Electrical Engineering faculty at Texas A&M University as a Professor of Electrical Engineering and Director of the Institute for Solid State Electronics in November 1985. Since 1988 he has held the Irma Runyon Chair in Electrical Engineering. During the summer of 1990, he was a Visiting Scientist at Nippon Telephone and Telegraph's Opto-Electronics Laboratory in Ibaraki Prefecture, Japan.

Since 1970 his principal research interests have been in the fields of fiber optics, integrated optics, and diode laser applications. He has authored more than 190 journal articles and conference presentations and holds 27 U.S. patents. He was awarded a Civil Service Commission/Navy fellowship to study Systems Analysis at the Massachusetts Institute of Technology during 1971–1972. He also received the Naval Electronics Laboratory Center Annual Science Achievement Award in 1974, the American Society of Naval Engineers' Solberg Award for Applied Research in 1975, and the Texas A&M Association of Former Students Award for Excellence in Research in 1991.

He served as Conference Chairman for the IEEE/Optical Society of America (OSA) Topical Meeting on Integrated and Guided Wave Optics for 1986, and was Program Chairman for that Conference in 1984. He was a member of the steering committee for the Optical Fiber Communication and Integrated Optics and Optical Communication Conferences from 1987 to 1989. He was Guest Editor for a Special Issue of the IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS, December 1979 and for a Special Issue of the *Journal of Lightwave Technology* for March 1987.

Dr. Taylor is a Fellow of OSA, a Life Member of the American Society of Naval Engineers, and a Member of the American Physical Society.