that problems due to deep traps have been dealt with in another paper [5]. A comparison with the silicon LDMOS shows that the silicon carbide impedances are greater than the silicon impedances, and it makes impedance matching easier for SiC. The forthcoming technological improvements will allow designers to build more powerful devices (we expect 100 W quite soon), and will make silicon carbide a good candidate for RF transistors in the next few years.

REFERENCES

1. A.W. Morse, P.M. Esker, S. Siriam, and R.C. Clarke, Recent application of silicon carbide to high power microwave, IEEE MTT-S Dig, 1997, pp. 53–56.
2. S.T. Allen et al., Silicon carbide MESFET's for high power S-band applications, IEEE MTT-S Dig, 1997, pp. 57–60.
3. O. Noblanc, C. Arnodo, E. Chartier, and C. Brylinski, Microwave power MESFET on 4H-SiC, Diamond and Related Mater, 1997, pp. 1508–1511.
4. T.T. Ha, Solid state microwave amplifier design, Wiley, New York, 1981.
5. O. Noblanc et al., Progress in the use 4H-SiC of semi-insulating wafers for microwave power MESFETs, ESCRM-98 Proc..

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A SIMPLE LOW-COST COPLANAR TWIN-SLOT ACTIVE ANTENNA OSCILLATOR FOR WIRELESS APPLICATIONS

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ABSTRACT: *This paper presents a compact, simple, low-cost coplanar waveguide active antenna oscillator which can be used for wireless communication applications. The circuit uses the active MGA86576 MMIC amplifier, which is biased through its output port using simple lumped elements. A closed feedback loop is implemented with two slot antennas fed by coplanar waveguide transmission lines. The paper presents a design procedure to systematically dimension the different parts of the structure until the oscillation condition is satisfied. An active antenna oscillator has been designed, manufactured, and tested, and measured results are in good agreement with predictions. The circuit exhibits a very clean, stable oscillation, and the radiation patterns are smooth and quasiomnidirectional. The measured gain of the active antenna is 3.2 dBi.* © 1999 John Wiley & Sons, Inc. Microwave Opt Technol Lett 23: 18–25, 1999.

Key words: *active antennas; oscillators; TRL calibration; slot antennas*

1. INTRODUCTION

In recent years, cellular and indoor communications have evolved very rapidly, and new RF circuits are constantly being investigated in order to answer the increasing demand of new communication systems in the market [1]. In this sense, the trend is to go to higher frequencies, where the frequency spectrum is less busy, and the transmission bandwidth naturally increases [2, 3]. In this case, the use of quasioptical frequency conversion structures is very convenient, since the received signal is mixed with the local oscillator just at the antenna terminals, so that losses in the circuits and undesired couplings are minimized [3, 4]. In many of these systems, the injection of the local oscillator signal is achieved through the same antenna, thus requiring active antenna oscillators which are able to illuminate all of the receivers. One such application is, for instance, in wireless LANs. In this case, an active antenna oscillator is placed in the room, with the task of illuminating the receivers placed on each computer. The oscillation signal thus injected is combined with the signal carrying the data, to produce the intermediate frequency (IF) at the receiver antenna terminals [5]. In addition, at high frequencies, the power of solid-state devices is limited, and to overcome this limitation, the active antenna oscillators are often designed together with power-combining techniques in order to increase the total power radiated [6].

Due to the proliferation of new communication applications, the investigation of active antenna oscillators has become very popular, and many configurations have been reported in the technical literature. For instance, microstrip patch antennas acting as active oscillators can be found in [4, 7]. The use of the coplanar waveguide for the design of active antenna oscillators is also very popular; see, for instance, [8–10]. The main advantage of this structure is that it is uniplanar, and therefore, manufacturing becomes very inexpensive. Also, integration with other active devices is simplified, mainly because via holes are eliminated for the connections to ground [8]. In spite of the advantages of uniplanar configurations, complex multilayered structures have also been used for the design of active antenna oscillators. See, for instance, [11], where a microstrip patch slot-coupled antenna is used, or [10], where a slot antenna and a slot loop are used to excite a microstrip patch, broadside coupled. In spite of all of these efforts, the investigation of new, simple, and compact structures with better electrical performances still remains an interesting activity.

A traditional approach for the design of active antennas is to consider the radiating element as a resonant load impedance of the active device. In consequence, the impedance of the antenna is adjusted to drive the active device (usually an FET or an HEMT transistor) into its unstable region. The design is completed by adding the required stubs for impedance matching at the input [8–10]. In the present work, a different approach for the design of the circuit is followed, as will be described next.

In this contribution, we have designed a simple, low-cost active antenna oscillator in uniplanar technology. The basis for the design is two slot antennas fed by coplanar waveguide transmission lines (see Fig. 1), to which an active MMIC amplifier is coupled. The main characteristic of the circuit is that a whole MMIC amplifier is used as an active element, and not just a single transistor as is usually done. In consequence, the power transmitted by the circuit is expected to be greater as compared to traditional designs. The first consequence of doing this is that the device used as an active element is very stable, and it is, therefore, difficult to design a passive load which can drive it into an unstable region. Due to this fact, the design is now carried out by considering the whole circuit as an amplifier with a closed feedback loop, which is used to produce the needed oscillation condition [see Fig. 1(b)]. Following this approach, one of the two slot antennas in Figure 1(a) is actually radiating most of the power. The second slot antenna is used to couple some of the radiated energy back to the input of the amplifier, therefore

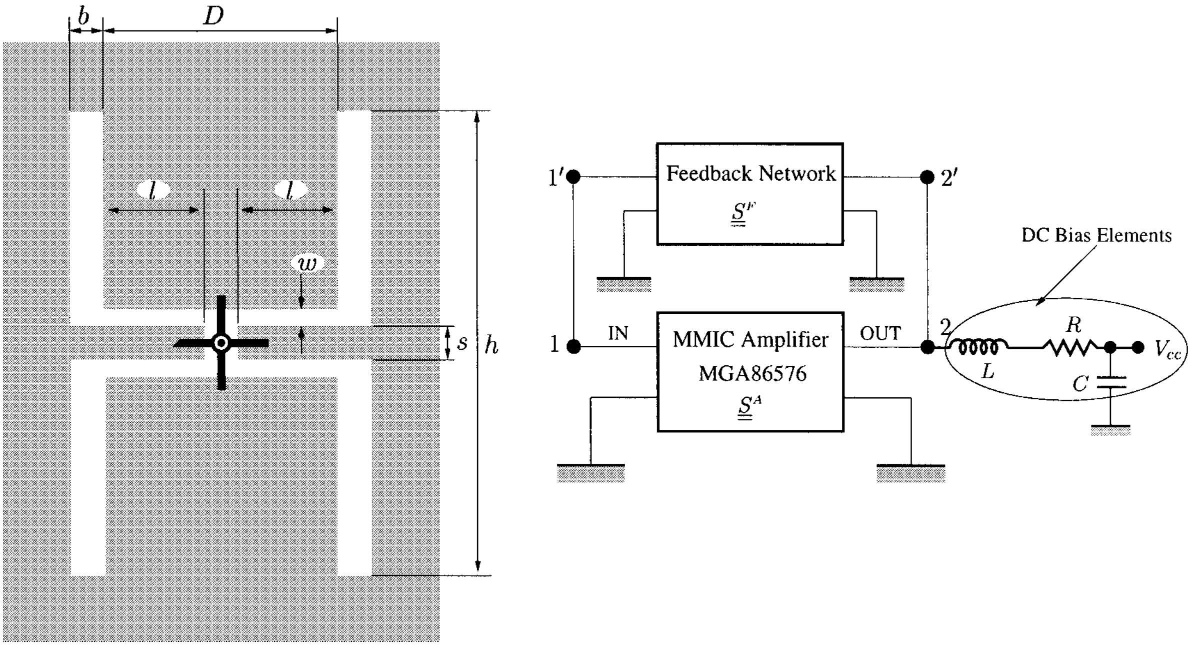


Figure 1 Basic coplanar waveguide active antenna oscillator geometry investigated in this paper

closing the feedback loop needed to establish the oscillation conditions.

The design is based on the scattering matrices of the active amplifier, and of the passive closed-loop feedback network. The scattering parameters of the amplifier are accurately measured through a TRL calibration, while a full-wave analysis technique is used for the calculation of the scattering parameters of the passive feedback network. Once they are known, the open-loop gain is computed, and oscillation is obtained with the usual conditions of total gain greater than 1 and total phase equal to 0°.

In this paper, first, the design procedure is described in detail. Then a prototype of the design circuit is manufactured and tested. The results are in very good agreement with the expected performance of the circuit. In particular, a very stable oscillation is obtained at the design frequency, with very smooth and quasiomnidirectional radiation patterns. Moreover, the gain of the active antenna is 3.2 dBi, which represents a good compromise between quasiomnidirectional patterns and good radiation efficiency.

2. DESIGN APPROACH

The active antenna oscillator investigated is presented in Figure 1(a). It is composed of an active element which, in the present case, is the inexpensive MMIC microwave amplifier MGA86576, and a passive network completing a feedback loop for the active device which, in the present design, is formed by coplanar waveguide transmission line sections feeding two coplanar slot antennas. The simplest equivalent circuit that can be used to ease the design of this structure is to consider the typical closed feedback loop network depicted in Figure 1(b). The reduction of the complete structure in Figure 1(a) to the equivalent network in Figure 1(b) is possible once the scattering parameters of both the active MMIC amplifier ($\underline{\underline{S}}^A$) and the passive feedback element ($\underline{\underline{S}}^F$) are known.

The analysis of the passive feedback network has been accomplished with a mixed-potential integral equation (MPIE) formulation using the multilayered media Green's functions developed in [12], and the slot elements are treated using equivalent magnetic currents in a similar way as addressed in [13]. The code developed allows us to obtain the full scattering parameters of a passive two-port coplanar circuit such as the one used as a feedback network in the active antenna (see Fig. 2), and therefore, it completely characterizes the first element of the oscillator. For our particular design, the whole feedback network shown in Figure 2 is printed on a TMM10 substrate of high relative permittivity ($\varepsilon_r = 9.2$) and thickness $h = 1.91$ mm. The dielectric substrate has been chosen relatively thick to ensure a minimum of rigidity in the final circuit, without the need for any additional support which might alter the electromagnetic behavior of the structure once assembled.

The last important element in our oscillator antenna is the active MMIC amplifier MGA86576. A fast and accurate way to obtain the scattering parameters of this device is through measurements, using the same configuration and biasing conditions that the component will have once mounted in the final oscillator circuit. Also, it is important to know the exact scattering parameters at the ports of the device, eliminating the effects of any connecting transmission line or input and output connectors. A simple way to do this is through the use of the so-called TRL calibration. Figure 3(a) shows the calibration kit in coplanar line technology, and Figure 3(b) presents the final test set where the MMIC is mounted for final measurements. The whole manufactured setup is suitable for measurement of devices in coplanar technology in the frequency band from 1 to 15 GHz. As seen in Figure 3(a), the calibration kit is composed of a thru line, a short-circuited line, and two coplanar waveguide transmission lines of different lengths. These two additional transmission lines are needed to correctly perform the calibration in two subbands (from 1 to 4 GHz and from 4 to 15 GHz), in which the whole frequency range has been divided for improving accuracy.

The final measurement of the active MMIC amplifier has been carried out on an HP8510C vector network analyzer,

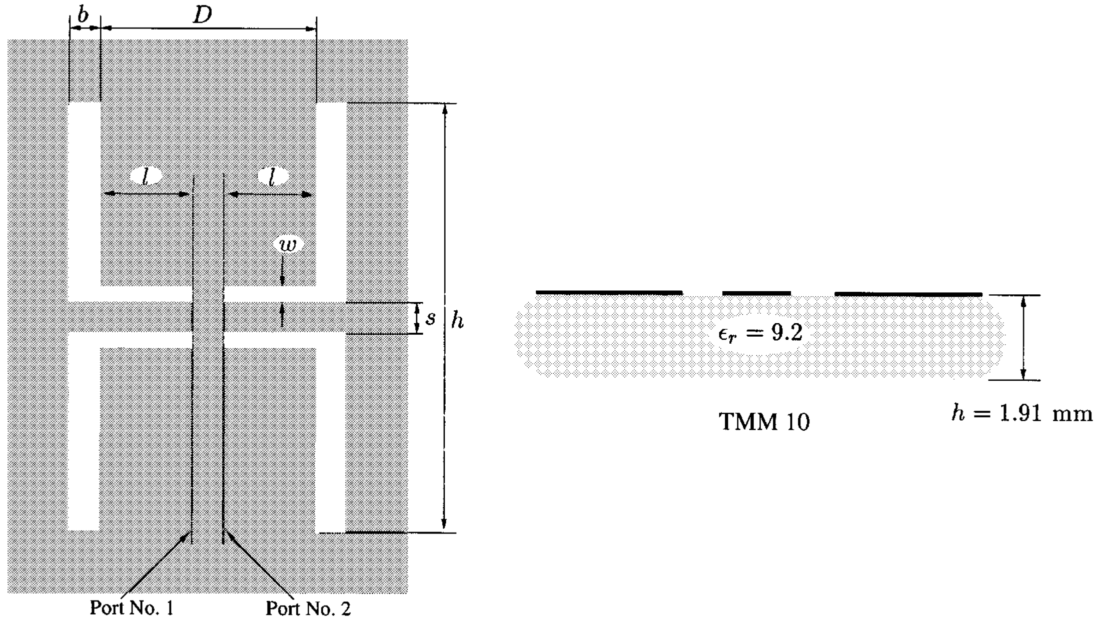


Figure 2 Basic geometry structure of the closed feedback loop, and the coplanar waveguide dielectric substrate used in the implementation

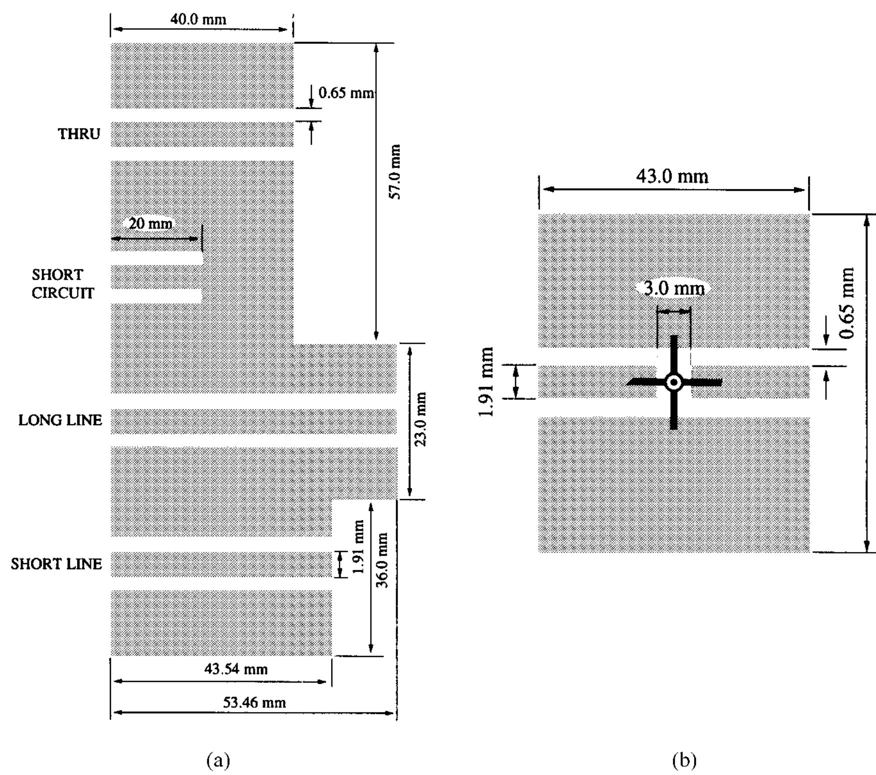


Figure 3 Geometry of the calibration kit in coplanar waveguide technology used to perform the TRL calibration, and the test set used for final measurements of the active MMIC MGA86576 amplifier

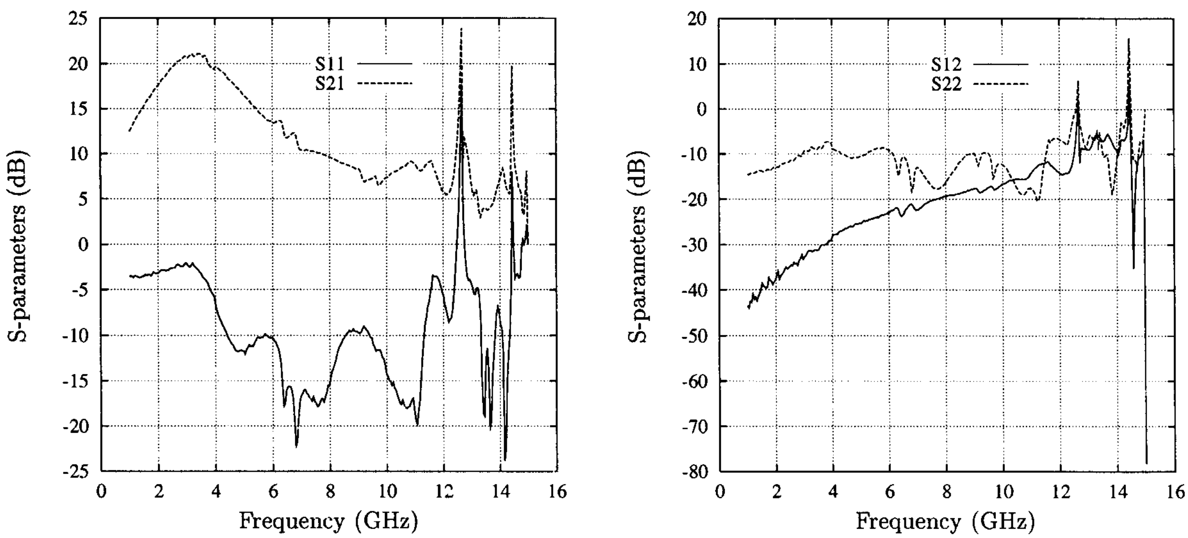
after successful TRL calibration using the manufactured kit. The MGA86576 is biased through its output port with a $V_{cc} = 6$ V dc power supply, adding a series $R = 56\ \Omega$ resistor, and a decoupling series inductance and parallel capacitance ($C = 100$ pF) used to isolate the RF signal from the dc circuitry.

The equivalent circuit of the whole active antenna including the bias circuit is shown in Figure 1(b). This setup provides the scattering parameters of the active MGA86576 measured at the connecting points of the device to the circuit. The same connecting pads and bias elements used during this measurement campaign will be used again in the design of the final active antenna in order not to modify the measured scattering parameters. Figure 4 presents the measured results, showing that about 20 dB of gain is obtained in the vicinity of the design frequency of 3.5 GHz.

At this point, it is interesting to mention that the scattering parameters of the MGA86576 amplifier have been measured under small-signal conditions. These are, therefore, the scattering parameters of the active device when the oscillation starts. However, once the oscillation is established, the active device will enter into saturation, and thus, the scattering parameters are those obtained under large-signal conditions, where the behavior is dominated by nonlinear interactions. If a single transistor is used, then the large-signal scattering parameters will be very different from those measured under small-signal conditions. However, for the MGA86576 MMIC amplifier, due to an internal feedback loop, the large-signal scattering parameters will not considerably differ from those measured under small-signal conditions, and in any case, they will be much more similar than in the case of a single transistor.

Once the scattering parameters of both the active device and the passive feedback network are known, the oscillation condition of the system can be checked out using the standard loop gain control theory. To do so, the closed feedback loop is first open, and then the gain of the resulting open loop is computed. The oscillation capability of the system can now be verified with the following two typical conditions, namely, gain of the loop greater than 1, and phase of the loop equal to 0°.

If the closed feedback loop of the equivalent circuit shown in Figure 1(b) is open, then the resulting network contains two cascaded quadripoles as depicted in Figure 5 ($\underline{\underline{S}}^A$, $\underline{\underline{S}}^F$).



(a) Excitation input port.

(b) Excitation output port.

Figure 4 Scattering parameters of the MMIC MGA86576 as measured with the HP8510C network analyzer with TRL calibration

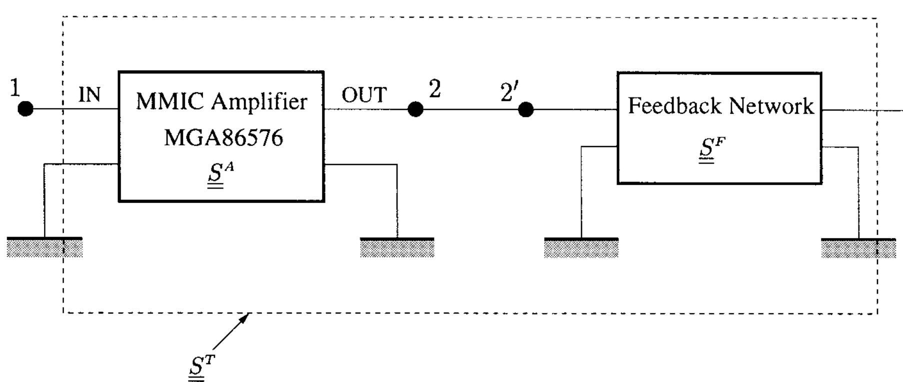


Figure 5 Open feedback loop equivalent network of the active antenna used for the study of the oscillation condition

In the present work, the gain of the feedback loop has been checked by computing the scattering parameters of the complete network in Figure 5, using the transmission chain matrix theory. First, the two individual scattering matrices are converted into transmission matrices. Next, the transmission matrices are multiplied to obtain the transmission matrix of the whole network. As a final step, the complete transmission matrix is transformed back to obtain the total scattering matrix of the ensemble ($\underline{\underline{S}}^T$).

It is interesting to note that the oscillation conditions can now be extracted from the total S_{21}^T-parameter thus calculated. In fact, the system will oscillate only if $|S_{21}^T| > 0$ dB, and the phase of S_{21}^T is strictly equal to zero. A software code has been built based on these principles, to allow for an easy optimization of the passive feedback loop until the oscillation conditions are satisfied. The next section presents the results of this investigation.

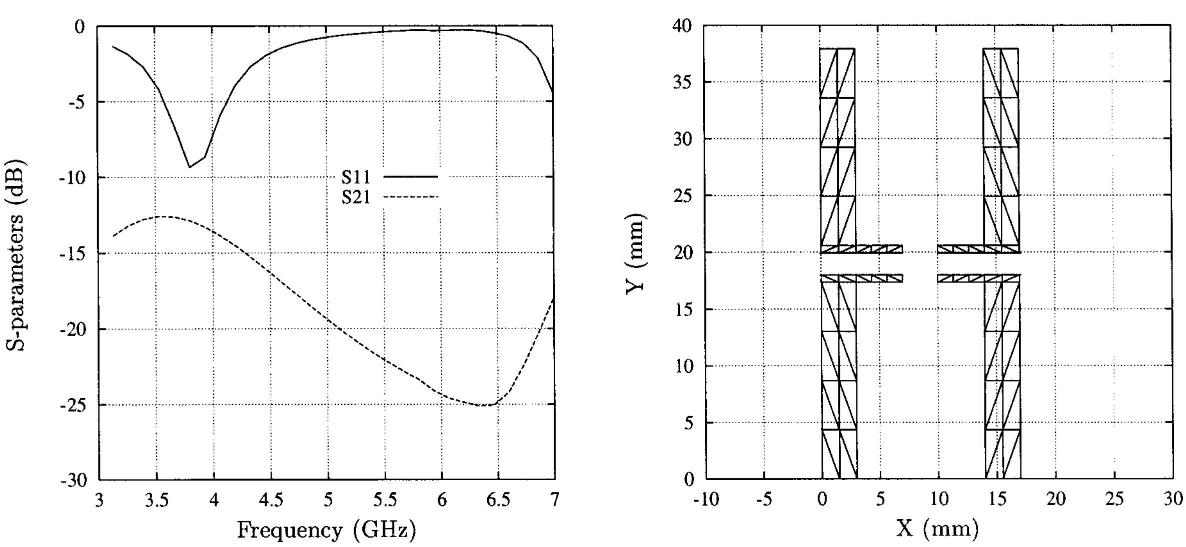
3. RESULTS

Using all of the tools developed in the previous section, an active antenna oscillator as shown in Figure 1(a) has been optimized at 3.5 GHz. From Figure 4, we observe that, at this frequency, the active MMIC amplifier exhibits a high gain of about 20 dB. Consequently, the first oscillation condition ($|S_{21}^T| > 0$ dB) is not very critical, and a coupling level between the two slot antennas greater than -20 dB will suffice. On the contrary, the condition for the phase of the S_{21}^T-parameter is very critical, and the feedback network shown in Figure 2 must be optimized so that the phase of the total S_{21}^T crosses the point of 0° at the desired oscillation frequency.

The scattering parameters of the feed network after optimization are presented in Figure 6. They correspond to the circuit shown in Figure 2 with $D = 11$ mm, $b = 3$ mm, $l = 4$ mm, $h = 37.9$ mm, and the coplanar waveguide transmission lines feeding the slot antennas are 50 Ω characteristic impedance lines, namely, $w = 0.65$ mm, $s = 1.91$ mm. As seen, there is a maximum of coupling between the two slot antennas around the design frequency of 3.5 GHz. The level of coupling is of -12 dB which, in principle, is enough to meet the first oscillation condition. The important point, however, is shown in Figure 7, where the phase of the total S_{21}^T-parameter is presented, indicating that there is a zero cross of the phase at the design frequency. Figure 7 also indicates that, at 3.5 GHz, both oscillation conditions are satisfied since the modulus of S_{21}^T is about 11 dB, and at the same time, the phase crosses the 0° value. Moreover, the figure also shows that no other spurious oscillation will occur in the band from 3 to 8 GHz. In fact, while the phase characteristic crosses the 0° point at two other frequencies, the gain of the loop is then lower than 1, so the oscillation condition is not satisfied at those frequencies. The spectrum of the active antenna is, therefore, expected to be clean, and thus, all design steps are completed.

The designed circuit has been manufactured, and Figure 8 shows the radiated spectrum of the active antenna measured with an HP8561E spectrum analyzer. The results indicate that the circuit is oscillating at the expected frequency of 3.5 GHz, with a very stable and neat emitted signal. In addition, Figure 8(b) shows that there are no spurious oscillations in the whole measured spectrum range from 100 MHz up to 6.5 GHz. Also, Figure 8(c) shows the second and third harmonics of the active antenna, indicating very good performance of the circuit. In fact, as can be seen, the level of the second harmonic is only -32.3 dBc, while the level of the third harmonic is -27 dBc, and they are separated in frequency a distance of about $\Delta f = 3.5$ GHz. This measurement again confirms that the active antenna is radiating a very clean spectrum, with very low power levels going to higher order harmonics.

Furthermore, Figure 9 presents the measured co- and cross-polar components of the radiated field, in both the E- and H-planes of the antenna. It can be observed that the radiation patterns are smooth, and they are also very omnidirectional. These radiation characteristics are very desirable in many cellular and indoor communication systems. It is also interesting to notice that the maximum gain of the antenna in the E-plane occurs at about 45 and 135°. The radiation in this plane is not, therefore, broadside. This is due to the fact that the structure is not symmetrical in this plane, mainly because one of the slot antennas is radiating most of the power.



(a) (b)

Figure 6 Simulated scattering parameters of the passive feedback network shown in Figure 2 after optimization

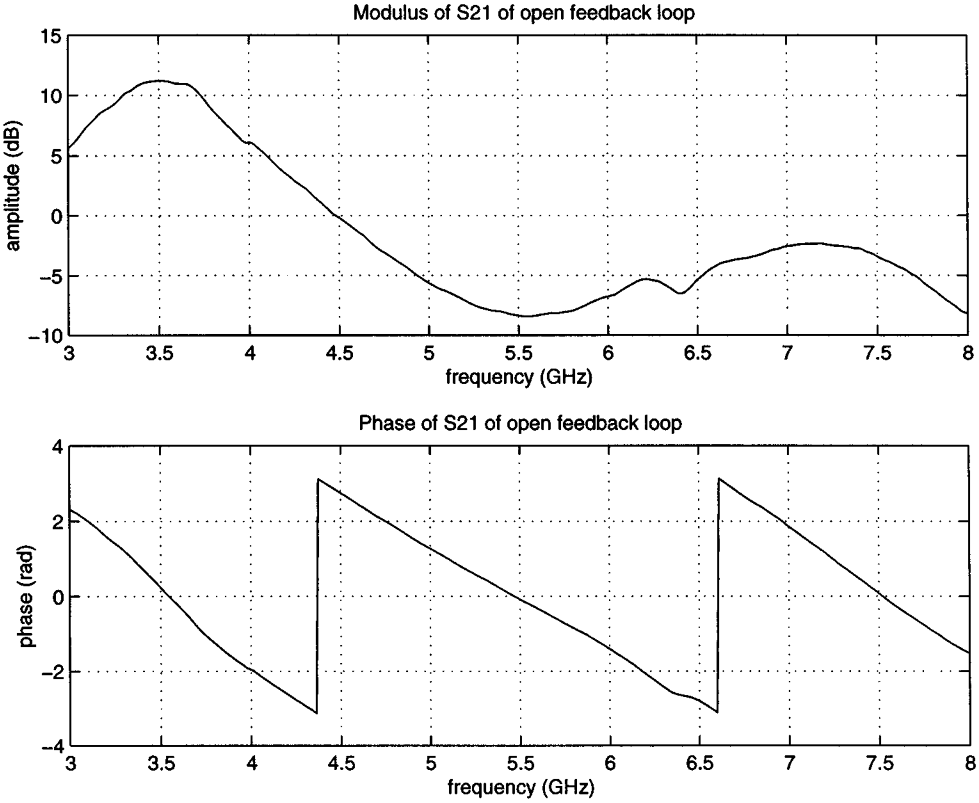


Figure 7 Scattering parameter S_{21}^{T} of the total open feedback network depicted in Figure 5, showing the oscillation condition fulfilled at 3.5 GHz

Indeed, the other slot antenna is simply used to receive some of the radiated energy, which is then delivered to the input port of the amplifier, as already mentioned.

In addition, Figure 10 presents the sensitivity of the circuit to a change in the dc bias voltage. The figure shows the frequency shift of the transmitted signal when the dc bias voltage is varied from the nominal 6 V initial value to 4.5 V. The results reveal a frequency shift of about 7 MHz when the bias voltage is changed. This is due to the fact that the scattering parameters of the amplifier change with the applied dc voltage. The figure also shows that the power transmitted decreases with the dc voltage, resulting from a reduction in the amplifier gain. This reduction in gain is not critical to maintain the oscillation, due to the margin of 11 dB obtained in the whole gain loop. The frequency shift, however, is mainly caused by a variation in the phase of the scattering parameters of the amplifier. This change results in the phase of the total S_{21}^{T}-parameter of the complete oscillator loop crossing the 0° point at a different frequency, therefore producing the observed shift in the oscillation condition of the active antenna.

The final measured data are the gain and the power transmitted by the active antenna oscillator. The circuit has been placed in an anechoic chamber with the HP8561E spectrum analyzer connected to the receiving antenna. First, the power transmitted by the active antenna, and received by the spectrum analyzer, is stored in memory. Then a reference generator connected to a reference antenna is placed in the position of the antenna oscillator, and the power of the generator is increased until the same power level is measured. The power of the reference generator which produces the same received signal level as the antenna oscillator is $P_{\mathrm{ref}} = 9.5$ dBm. In addition, the gain of the reference antenna, connected to the reference generator, is $G_{\mathrm{ref}} = 2.1$ dBi. With all of this information, the product gain times the power emitted by the active oscillator can be obtained through the relation

$$P_{\mathrm{ref}} G_{\mathrm{ref}} = P_{\mathrm{osc}} G_{\mathrm{osc}} \tag{1}$$

where P_{osc}, G_{osc} are the power and gain of the active antenna oscillator, respectively. The power P_{osc} can be considered to be the maximum power that can be delivered by the active MGA86576 amplifier. Measurement of this maximum power gives a value of $P_{\mathrm{osc}} = 8.4$ dBm. The gain of the active antenna is then computed with Eq. (1), obtaining $G_{\mathrm{osc}} = 3.2$ dBi. This value is a good compromise between a radiation pattern not too directive, still keeping a good radiation efficiency of the active antenna oscillator.

4. CONCLUSIONS

This paper has presented a simple active antenna oscillator in coplanar waveguide technology, built using the inexpensive MGA86576 MMIC amplifier, to which a passive feedback network is coupled. The feedback network is uniplanar, and it is composed of two coplanar waveguide transmission lines feeding two slot antennas.

In this paper, the design procedure leading to the dimensioning of the whole structure is carefully reviewed. The method is based on extracting the scattering matrices of both the active MMIC amplifier and the passive feedback network. For the active device, the scattering parameters are measured using a TRL calibration process. For the passive feedback

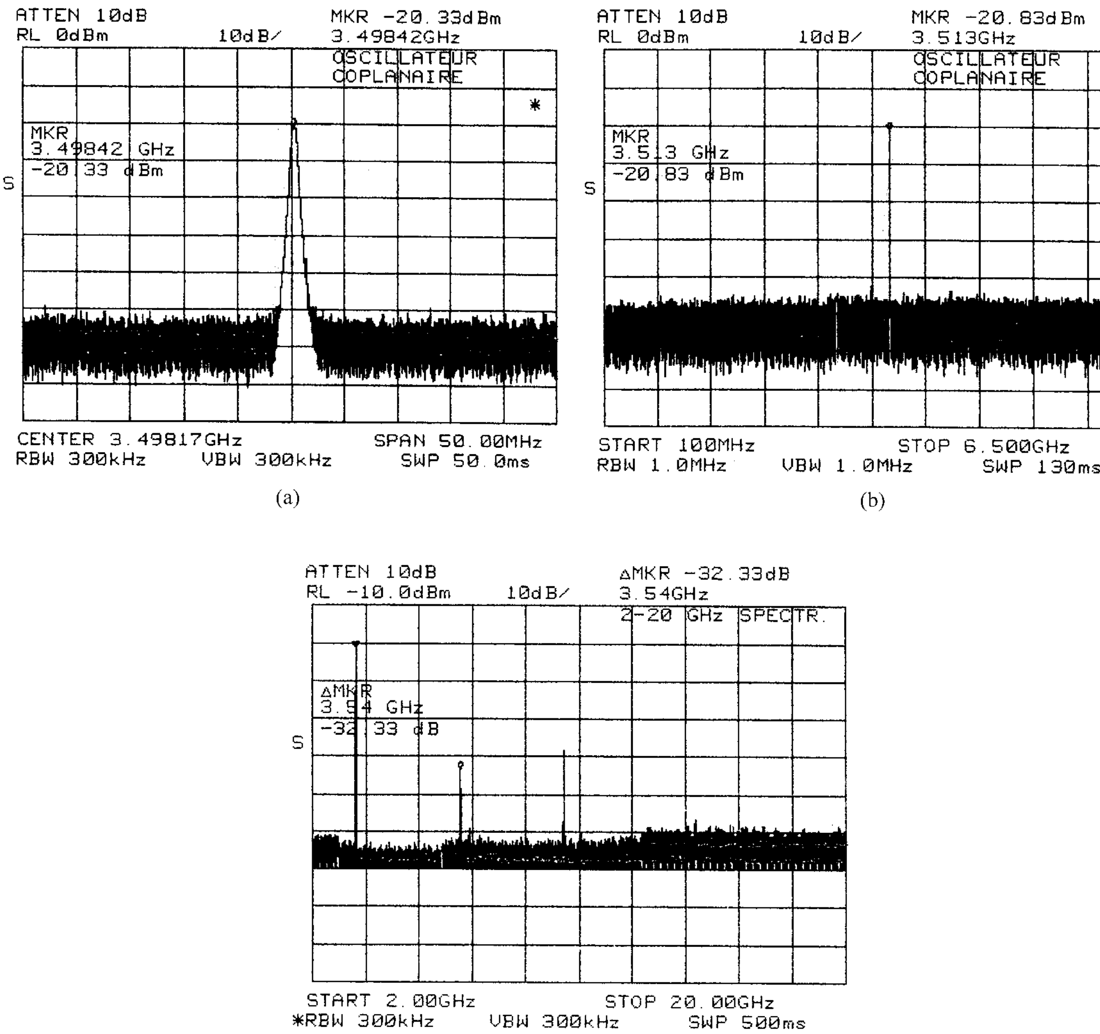


Figure 8 Frequency spectrum transmitted by the manufactured active antenna oscillator

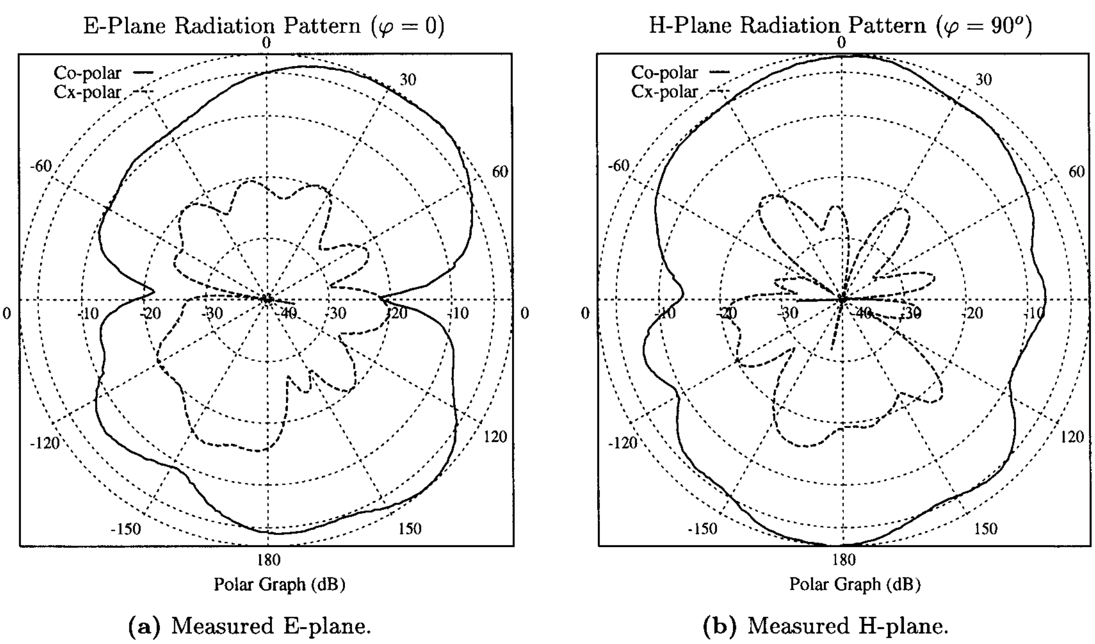


Figure 9 Measured far-field radiation patterns of the active antenna oscillator

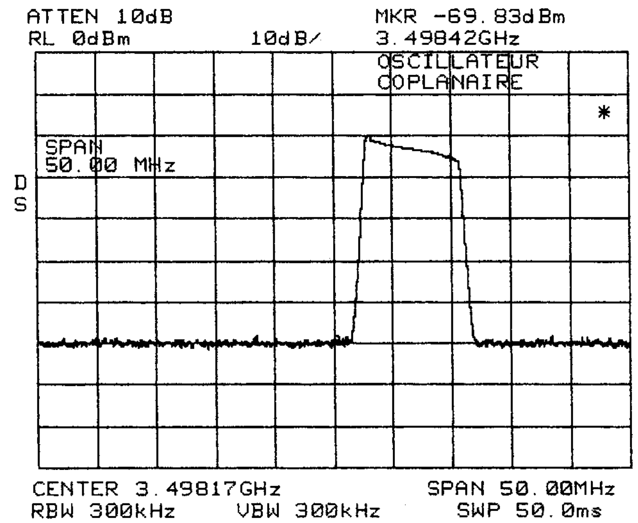


Figure 10 Frequency drift as a function of the dc voltage applied to the active antenna oscillator. Voltage is varied from nominal $V_{cc} = 6$ V to $V_{cc} = 4.5$ V

network, a standard MPIE technique using the multilayered media Green's functions is implemented.

Once the scattering matrices of both components are obtained, the loop gain is computed by evaluating the total scattering matrix of the open feedback loop network. A software code has been implemented to check the oscillation condition.

After the design process, an optimized structure has been manufactured and tested. Results show that the active antenna is correctly operating at the design frequency. Moreover, the oscillation obtained is very stable, and no spurious signals (clean spectrum) are obtained. Moreover, the far-field radiation shows clean, symmetric radiation patterns. Finally, the power level radiated is good, and the measured gain of the antenna is about 3.2 dBi, which represents a good compromise between quasiomnidirectional patterns and good radiation efficiency.

REFERENCES

1. R.B. Waterhouse and D. Novak, Low cost printed antenna suitable for indoor wireless base stations, Proc IEEE AP-S Symp, Microwave Antennas and Propagat, Baltimore, MD, July 1996, vol. 1, pp. 46–49.
2. C.C. Ling and G.M. Rebeiz, A 94 GHz planar monopulse tracking receiver, IEEE Trans Microwave Theory Tech 42 (1994), 1863–1871.
3. K.D. Stephen, N. Camilleri, and T. Itoh, A quasi-optical polarization duplexed balanced mixer for millimeter-wave applications, IEEE Trans Microwave Theory Tech MTT-31 (1983), 164–170.
4. J. Lin and T. Itoh, Active integrated antennas, IEEE Trans Microwave Theory Tech 42 (1994), 2186–2194.
5. P. Otero, A. Alvarez-Melcon, J.F. Zürcher, and J.R. Mosig, A microstrip-coupled slot-loop antenna for integrated receivers in the millimeter-wave band, Microwave Opt Technol Lett 18 (1998), 91–95.
6. J.W. Mink, Quasi-optical power combining of solid state millimeter wave sources, IEEE Trans Microwave Theory Tech MTT-34 (1986), 273–279.
7. V.B. Erturk, R.G. Rojas, and P. Roblin, Design, analysis of an active integrated antenna, Proc IEEE AP-S Symp, Microwave Antennas and Propagat, Baltimore, MD, July 1996, vol. 2, pp. 1322–1325.
8. B.K. Kormanyos, W. Harokopus, L.P.B. Katehi, and G.M. Rebeiz, CPW-fed active slot antennas, IEEE Trans Microwave Theory Tech 42 (1994), 541–545.
9. G. Forma and J.M. Laheurte, Compact oscillating slot loop antenna with conductor backing, Electron Lett 32 (1996), 1633–1635.
10. G. Forma and J.M. Laheurte, CPW-fed oscillating microstrip antennas, Electron Lett 32 (1996), 85–86.
11. J.F. Zürcher and Y. Brand, An active strip-slot-foam inverted patch antenna, Microwave Opt Technol Lett 13 (1996), 114–119.
12. J.R. Mosig, "Integral equation technique," Numerical techniques for microwave and millimeter-wave passive structures, T. Itoh (Editor), Wiley Interscience, New York, 1989, Chap. 3, pp. 133–213.
13. R.C. Hall and J.R. Mosig, The analysis of arbitrarily shaped aperture-coupled patch antennas via a mixed-potential integral equation, IEEE Trans Antennas Propagat 44 (1996), 608–614.

LINEAR MICROSTRIP ARRAY ANTENNA WITH A SIMPLE FEED NETWORK

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ABSTRACT: *A simple feed network is introduced for a linear microstrip array antenna. The antenna structure consists of two layers. The radiating microstrip patches are placed on the top surface of the upper layer, while the bottom layer is used to feed the radiating elements through the slots on the conducting microstrip line between the upper and lower layers. The resultant array antenna is relatively simple, and the feed loss can be minimized.* © 1999 John Wiley & Sons, Inc. Microwave Opt Technol Lett 23: 25–27, 1999.

Key words: *array antenna; microstrip antenna*

I. INTRODUCTION

In a recent paper by one of the authors (Lee) [1], a simple double-layer microstrip array antenna was introduced. The top layer of the antenna contains radiating microstrip patches. The bottom layer is a transmission line with slots under the radiating patches. As the electromagnetic energy propagates along the lower layer, part of the energy leaks at each slot, and the far fields from the radiating patches interfere constructively to produce a focused beam. The device is simple and inexpensive to fabricate. However, since the antenna operation is based on a traveling wave, all of the remaining energy in the transmission line after the last radiating patch will be wasted, resulting in low antenna efficiency. This paper introduces an array antenna that is similar to the previous design, but that uses a standing wave to improve the antenna efficiency.

In Section II, the design scheme is described, followed by the theoretical and experimental results in Section III.